OCTOBER · 1954

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Journal of Communications and Electronic Engineering

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Volume 42 Number 10

OVEL
UBE
ESIGN

Pictured above with its ceramic envelope removed is a "stacked tube," a new development in which tube elements are simply stacked one on top of the other with ceramic spacers between. This novel design results in a compact (see left) and highly rugged device which readily lends itself to mass production techniques. Left, comparative photo of conventional model and new "stacked tube."

Sylvania Electric Products Inc.

1955 I·R·E CONVENTION AUTHORS
See page 1573 of this issue.

IN THIS ISSUE

- Beam Deflection Color TV Picture Tubes
- Viewing Storage Tube with Halftone Display
- IRE Standards on Semiconductor Terms
- Impedance of Periodically Loaded Structures
- Microwave Correlator
- Calibration of AM Meters
- A New Transducer Diagram
- Making a Radar Self-Calibrating
- Laminated Conductor for High Frequencies
- Transmission Line Directional Couplers
- Power-Spectra of Noise-Modulated Carrier
- Discontinuities in Disk Transmission Lines
- Positive-Ion Trapping in Electron Beams
- Impedance Measurement by Means of a Coupler
- Weighting Functions for Feedback Systems
- Propagation Modes in Dielectric Sheets
- *Transactions* Abstracts
- Abstracts and References

TABLE OF CONTENTS INDICATED BY BLACK-AND-WHITE MARGIN, FOLLOWS PAGE 80A

IRE Standards on Definitions of Semiconductor Terms appear in this issue.

The Institute of Radio Engineers

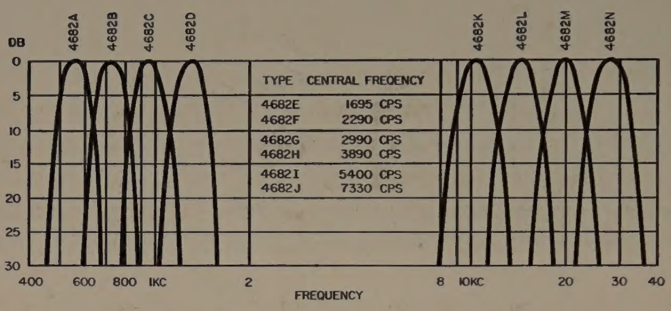
OUR MILLIONTH FILTER SHIPPED THIS YEAR...

FILTERS

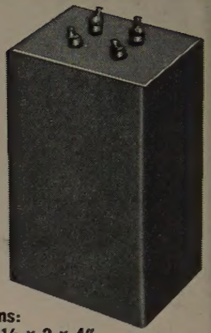
FOR EVERY APPLICATION

TELEMETERING FILTERS

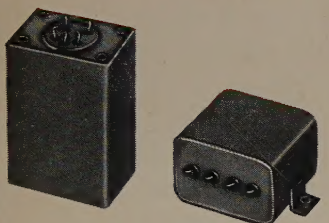
UTC manufactures a wide variety of band pass filters for multi-channel telemetering. Illustrated are a group of filters supplied for 400 cycle to 40 KC service. Miniaturized units have been made for many applications. For example a group of 4 cubic inch units which provide 50 channels between 4 KC and 100 KC.



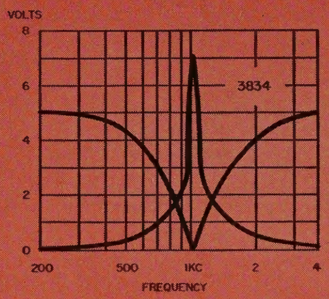
TYPE	CENTRAL FREQENCY
4682E	1695 CPS
4682F	2290 CPS
4682G	2990 CPS
4682H	3890 CPS
4682I	5400 CPS
4682J	7330 CPS

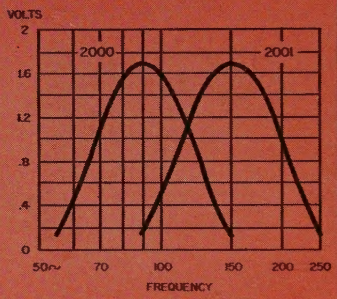


Dimensions:
(4682A) 1½ x 2 x 4"



Dimensions:
(3834) 1¼ x 1¾ x 2-3/16".
(2000, 1) 1¼ x 1¾ x 1⅝".





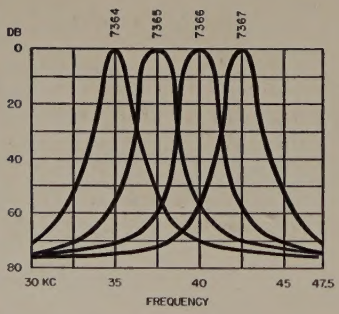
AIRCRAFT FILTERS

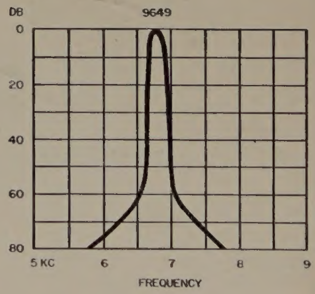
UTC has produced the bulk of filt
used in aircraft equipment for ov
a decade. The curve at the left
that of a miniaturized (1020 cycl
range filter providing high attenu
tion between voice and range f
quencies.

Curves at the right are that of o
miniaturized 90 and 150 cycle filte
for glide path systems.

CARRIER FILTERS

A wide variety of carrier filters are available for specific applications. This type of tone channel filter can be supplied in a varied range of band widths and attenuations. The curves shown are typical units.

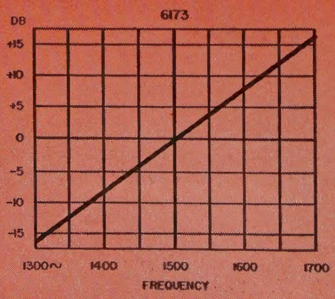


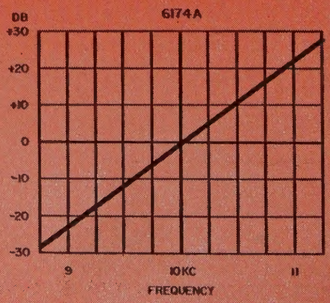


Dimensions:
(7364 series) 1⅝ x 1⅝ x 2¼".
(9649) 1½ x 2 x 4".

DISCRIMINATORS

These high Q discriminators provide exceptional amplification and linearity. Typical characteristics available are illustrated by the low and higher frequency curves shown.



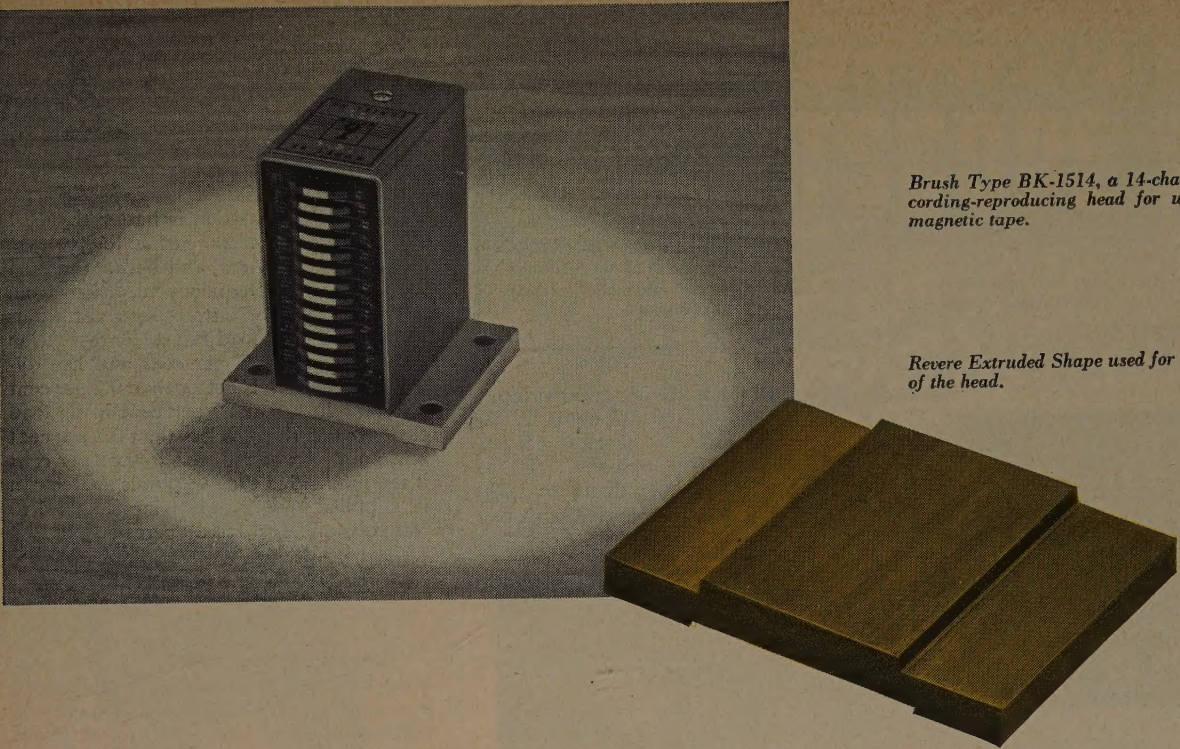


Dimensions:
(6173) 1-1/16 x 1⅜ x 3".
(6174A) 1 x 1¼ x 2¼".

For full data on stock UTC transformers, reactors, filters, and high Q coils, write for Catalog A.

UNITED TRANSFORMER CO

150 Varick Street, New York 13, N. Y. EXPORT DIVISION: 13 E. 40th St., New York 16, N. Y. CABLES: "ARLA



Brush Type BK-1514, a 14-channel recording-reproducing head for use with magnetic tape.

Revere Extruded Shape used for the base of the head.

Revere Extruded Shape

Saves 15¢ per piece

For Brush

The base of the magnetic recording-reproducing head shown here is a Revere brass extruded shape. You can see that it is rather simple in design, yet Brush Electronics Co., Cleveland, Ohio, reports that the shape saves 15¢ per piece (1½″ long) over the previous method of milling the piece out of solid bar. There were three operations required on the bar, which weighed 1.61 lb. per foot, against 1.22 lb. per foot for the shape. Eliminating the machining operations, and reducing scrap almost to the vanishing point, produced the economy.

The head in question can record and reproduce signals from 14 channels, at frequencies within, below, or above the audible range. Such a head is being increasingly employed to handle information to be used for computation, telemetering, inventory records, process control (automation) and similar purposes.

Extruded shapes by Revere should be looked into if you are doing any extensive machining of raw stock in copper and its alloys, and aluminum alloys. The extrusion process is much like squeezing paste from a tube. Much more intricate shapes than the one shown here are possible. Naturally, all design lines must be parallel to the axis of extrusion. Get in touch with the nearest Revere Sales Office, and see if Revere Extruded Shapes may not save you money.

REVERE
COPPER AND BRASS INCORPORATED
Founded by Paul Revere in 1801
230 Park Avenue, New York 17, N. Y.

• • •

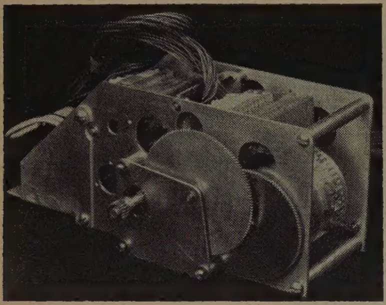
Mills: Baltimore, Md.; Chicago and Clinton, Ill.; Detroit, Mich.; Los Angeles and Riverside, Calif.; New Bedford, Mass.; Rome, N. Y.—
Sales Offices in Principal Cities, Distributors Everywhere.
SEE "MEET THE PRESS" ON NBC TELEVISION, SUNDAYS

NEWS and NEW PRODUCTS

October 1954

Digital Converter

A new digital converter, for digital data reduction and handling, has just been introduced by the recently-formed **Data Reduction and Automation Div., Fischer & Porter Co.,** 22 Jacksonville Rd., Hatboro, Pa.



The function of the Digi-Coder is to convert into a digital signal the analog output of primary sensing devices, such as flow meters, thermocouples or pressure transducers, and so forth. Digital signals, represented by discrete contact positions in the Digi-Coder, are the required input to electric typewriters, digital computers, punched card equipment, and other data-handling equipment.

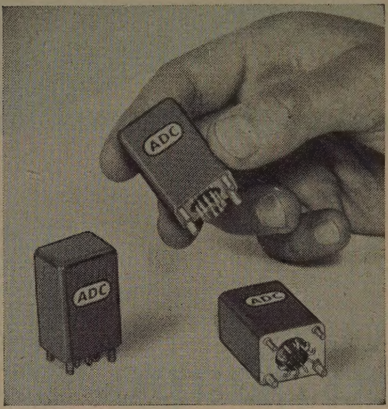
Two basic types of Digi-Coder are offered. One type converts a mechanical motion, delivered to its input shaft from a pneumatic or other receiving mechanism, into a digital output signal. Torque requirements are less than 0.20 inch/ounce. Low moment of inertia makes the converter adaptable to any servo instrument. The second type of Digi-Coder, which contains an integral self-balancing potentiometer, converts an analog voltage input into the digital signal. The voltage-input type will operate on a 0–1 millivolt input and up, with either a linear or nonlinear relationship maintained between input and output signals.

The Digi-Coder is accurate to the nearest digit, regardless of the number of digits in the output. In a 5-drum train, demonstrated accuracy is better than 1 in 34 billion parts. The Digi-Coder may be obtained with a hysteresis-free input coupler than enables the feelers to operate without interrupting the motion of the input shaft. The device can operate at speeds up to 5,000 counts per second.

Miniature Transformers

A recent announcement by the **Audio Development Co.,** 2833 13 Ave. S., Minneapolis, Minn., covers a new line of miniature hermetically-sealed transformers and chokes. Suitable for use in circuitry such as geophysical or transistor application, these compact transformers and chokes measure $\frac{3}{4}\times\frac{15}{16}\times1\frac{7}{8}$ inches. Available in both standard steel and nu metal, they are classified into the following groups: input 500 (125) ohms to 150,000 ohms C. T., primary inductance 8.5 henries ±5 per cent; output, 20,000 ohms C. T. to 16/8/4/2 ohms, primary inductance 1,200 henries; inductors, 1,500 henries, with two 2 per cent taps.

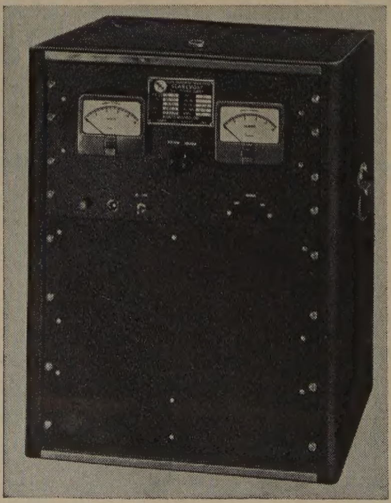
These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.



Special designs are also available up to 10-watt, 400-cps power transformers, audio transformers and inductors with ratings similar to those shown above. For additional information address your inquiry to: Industrial Sales Div., Audio Development Co.

Computer Power Supply

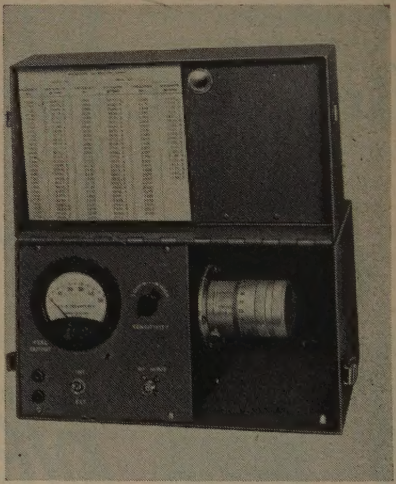
Magnetic Research Corp., 318 Kansas St., El Segundo, Calif., has developed a magnetic regulated dc power source for digital and analog computers, called the stabevolt type C. It is doubly magnetically regulated and has low internal impedance and close dynamic regulation.



Dynamic regulation is better than 0.1 per cent for step change of ac line voltage from 95 to 135 volts, and better than 0.1 per cent for line frequency variations from 57 to 63 cps. For output load variations from zero to full load the static regulation is better than 0.15 per cent and the dynamic regulation is better than 0.2 per cent for step change of 20 per cent in dc load current. The ripple is less than 0.2 per cent RMS and can be reduced to 0.01 per cent if required. This type of power source can be supplied in ratings ranging from 1.5 to 500 volts and 1 to 100 amperes output.

Field-Test Frequency Meters

Frequency Standards, Box 504, Asbury Park, N. J., has recently announced a new line of field test frequency meters in addition to their standard line of wavemeters. The ranges covered by these instruments are: 900 to 1,200 mc, 1,200 to 1,700 mc, 1,700 to 2,300 mc, 4,400 to 5,800 mc, 5,800 to 8,200 mc.



Accuracy of models 1217 and 1723 is 0.02 per cent; in all others it is 0.01 per cent.

A usable indication on the self-contained micro-ammeter is obtained with 1 milliwatt input.

All models have type BNC input and accessories such as coax-to-waveguide transducers and cables, which are available for the particular requirements of the customer. Each meter is individually calibrated and charts or curves are supplied. The frequency-determining elements of the wavemeters are made of invar to keep the temperature coefficient of the instruments as low as possible.

Models Nos. 4458 and 5882, which operate in the TE_{111} mode, are sealed and feature a built-in desiccator to eliminate the effect of varying relative humidity on the calibration of the instrument. Readers are invited to request new bulletins from the firm.

(Continued on page 27A)

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 20A)

Analog-to-Digital Converter

An analog-to-digital converter especially designed for use in data handling systems or digital computers has been announced by **Librascope, Inc.,** 1607 Flower St., Glendale, Calif.

ANALOG-DIGITAL CONVERTER
LIBRASCOPE
MANUFACTURED BY LIBRASCOPE INCORPORATED GLENDALE CALIFORNIA

Signal travels from shaft position to code discs and thence to either a relay system or a diode scanning network followed by a single flip-flop. Time sharing of control circuits is possible for multi channel systems using isolation diodes in series with each pick-off brush, allowing control circuit to be used for scanning many converters.

Ambiguity is eliminated by using two pick-off brushes for each track of the code disc except the first, or least significant. Standard code discs are manufactured for binary operation, but Gray, decimal, or other systems may be ordered. Present models include 7, 13, 17, and 19 digit units, all 2 inches in diameter and varying in length from 2 49/64 to 4 13/16 inches.

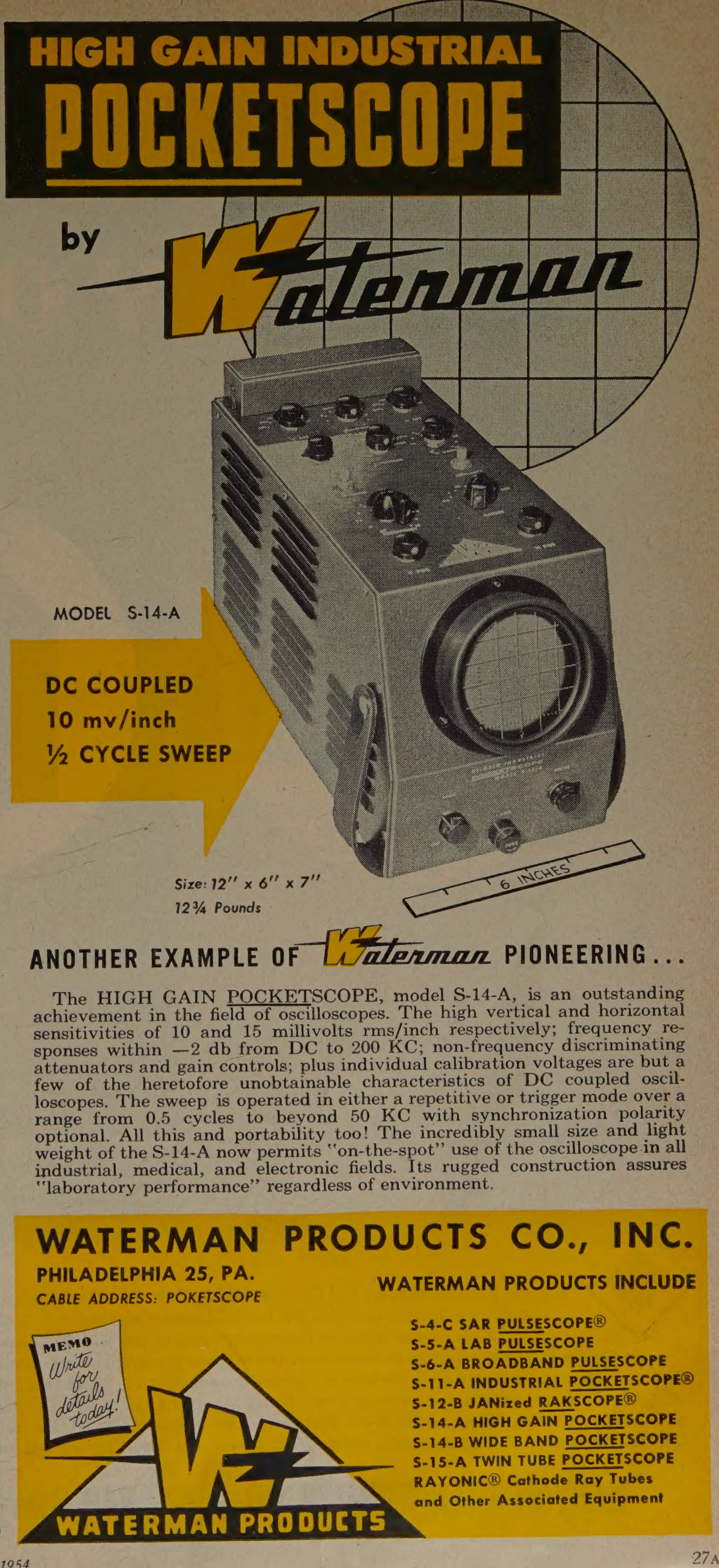
Conversion from digital code to analog shaft rotation may be made, but with a slight increase in circuitry.

Specifications: Input torque, under 0.2 ounce/inch. Current carrying capacity: 2 ma per pick-off brush. Digit output frequency may vary from 500 cps to 1 mc or higher. Stability: Operates independent of voltage, temperature, frequency limitations over a wide range. Life Expectancy: 5 Million or more cyclings of input shaft at speeds of 1 to 2 RPS. Weight: 8 ounces.

Color Bar White Dot Generator

A new color television test instrument, designed to produce color bars, white dots, or a crosshatch pattern, has just been an-

(Continued on page 32A)



HIGH GAIN INDUSTRIAL
POCKETSCOPE
by Waterman

MODEL S-14-A

DC COUPLED
10 mv/inch
½ CYCLE SWEEP

Size: 12″ x 6″ x 7″
12¾ Pounds

ANOTHER EXAMPLE OF Waterman PIONEERING...

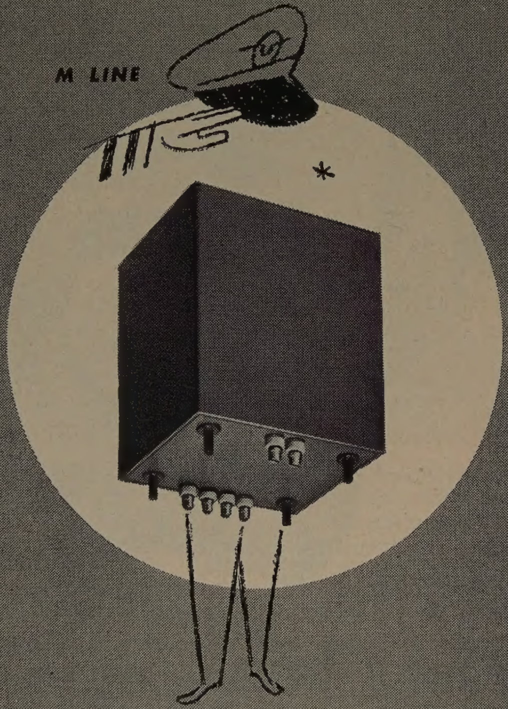
The HIGH GAIN POCKETSCOPE, model S-14-A, is an outstanding achievement in the field of oscilloscopes. The high vertical and horizontal sensitivities of 10 and 15 millivolts rms/inch respectively; frequency responses within −2 db from DC to 200 KC; non-frequency discriminating attenuators and gain controls; plus individual calibration voltages are but a few of the heretofore unobtainable characteristics of DC coupled oscilloscopes. The sweep is operated in either a repetitive or trigger mode over a range from 0.5 cycles to beyond 50 KC with synchronization polarity optional. All this and portability too! The incredibly small size and light weight of the S-14-A now permits "on-the-spot" use of the oscilloscope in all industrial, medical, and electronic fields. Its rugged construction assures "laboratory performance" regardless of environment.

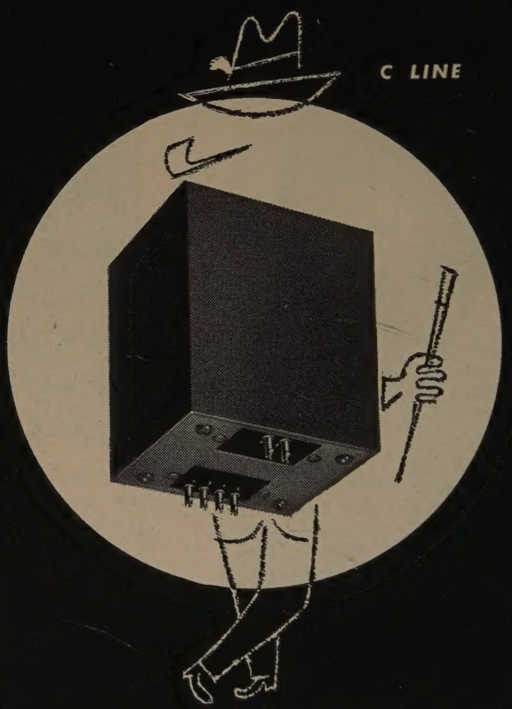
WATERMAN PRODUCTS CO., INC.
PHILADELPHIA 25, PA.
CABLE ADDRESS: POKETSCOPE

WATERMAN PRODUCTS INCLUDE
S-4-C SAR PULSESCOPE®
S-5-A LAB PULSESCOPE
S-6-A BROADBAND PULSESCOPE
S-11-A INDUSTRIAL POCKETSCOPE®
S-12-B JANized RAKSCOPE®
S-14-A HIGH GAIN POCKETSCOPE
S-14-B WIDE BAND POCKETSCOPE
S-15-A TWIN TUBE POCKETSCOPE
RAYONIC® Cathode Ray Tubes
and Other Associated Equipment

ELECTRICALLY AND MECHANICALLY INTERCHANGEABLE

THE KENYON TWINS





MEETS ALL MIL-T-27 REQUIREMENTS

Completely hermetically sealed to meet MIL-T-27 specifications. Military Standard types are included. Catalog Listing M L

MEETS ALL CIVILIAN REQUIREMENTS

A complete line to MIL electrical specifications, housed in conventional cases. Satisfactory performance at substantial savings. Catalog Listing C L

HOUSED IN CASES WITH STANDARD MIL-T-27 DIMENSIONS, BOTH LINES CARRY THE USUAL KENYON GUARANTEE

The new Kenyon Military and Commercial Lines feature the very latest practice, *using the best class "A" wire and insulation now available, and the latest types of core material,* to obtain minimum size at reasonable cost. . . . Cases are identical to the requirements in the MIL-T-27 specification and standard units are finished in a smooth, durable medium gray. Other colors are available on special order. Full rating information and schematic is furnished in the form of a stencil on each case. . . . Special units with other ratings and the same or similar cases are available on short delivery, in any quantity. Class B, class C, and class H units can also be delivered no special order. Where casing is not required yet sealing is important, we recommend our "Ken-Seal" molded transformers, available on special order. Write for catalog. Your inquiry will receive prompt attention.

KENYON TRANSFORMER CO., INC., 840 Barry St., New York 59

SPECIFIC TUBE DESIGN SERVICE • SPECIFIC TUBE APPLICATION SERVICE • LOCAL LABORATORY-TESTING SERVICE

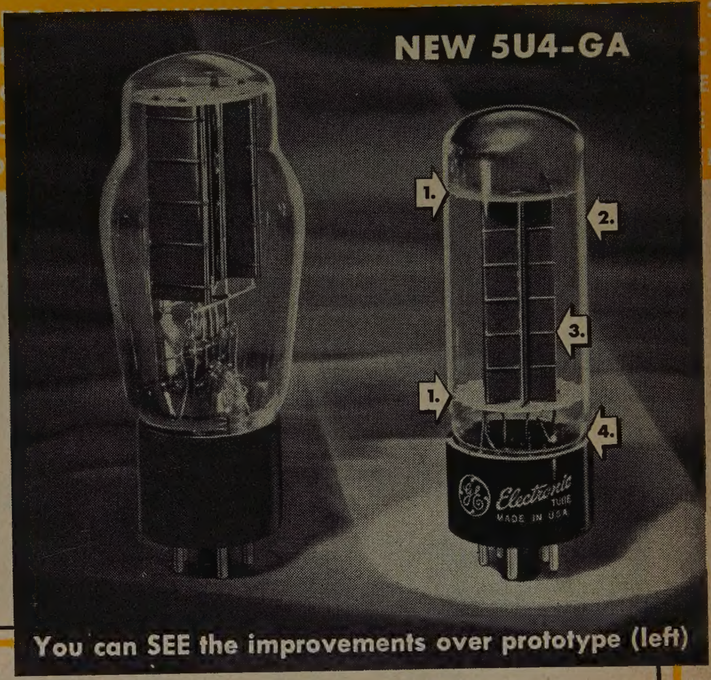
NON-STOP ORDER AND DELIVERY SERVICE • SPECIFIC TUBE DESIG

SPECIFIC TUBE APPLICATION SERVICE • LOCAL LABORATORY-TESTING SERVICE •

COAST-TO-COAST WAREHOUSING SERVICE • NON-STOP ORD

Specify NEW G-E LONG-LIFE RECTIFIER TUBES*!

✻ 5AU4
✻ 5U4-GA
✻ 5Y3-GT



You can SEE the improvements over prototype (left)

Sturdy, resist shocks, dissipate heat efficiently...yet prices are unchanged!

THE improvements you can SEE in the new 5U4-GA apply to all three new G-E rectifier tubes for television. G-E Design Service brings you, *at no price increase*, two rectifier tubes that are completely re-engineered, plus one brand-new type, the 5AU4. All are much more dependable than present types, so help reduce your TV production-line rejects. The new tubes are longer-lived—they cut down on service call-backs, increasing the reputation of your sets.

Recent introduction of new 6BQ6-GA and 25BQ6-GA sweep tubes, and this announcement of new G-E rectifier types, are only the first steps in an extensive General Electric program to design and build greatly improved receiving tubes for TV. Manufacturers of sets are asking for better tubes . . . G.E. is devoting every resource to the task of supplying them!

Keep in touch with G.E. for new-design tubes that will mean *new* high quality, *new* value, *new* reliable performance in the receivers you design and manufacture! Address *Tube Department, General Electric Company, Schenectady 5, New York.*

The 5U4-G prototype, though it did a good electrical job, was subject to damage from shocks and vibration. G.E.'s new 5U4-GA withstands hard usage, gives long service. Arrows (above, right) point to reasons why:

1.

Substantial mica supports brace the tube structure at both top and bottom, instead of at the top only.

2.

Glass bulb now is straight-side, compact, and strong. Diameter is 30% less than 5U4-G.

3.

New double-fin plate construction improves heat dissipation.

4.

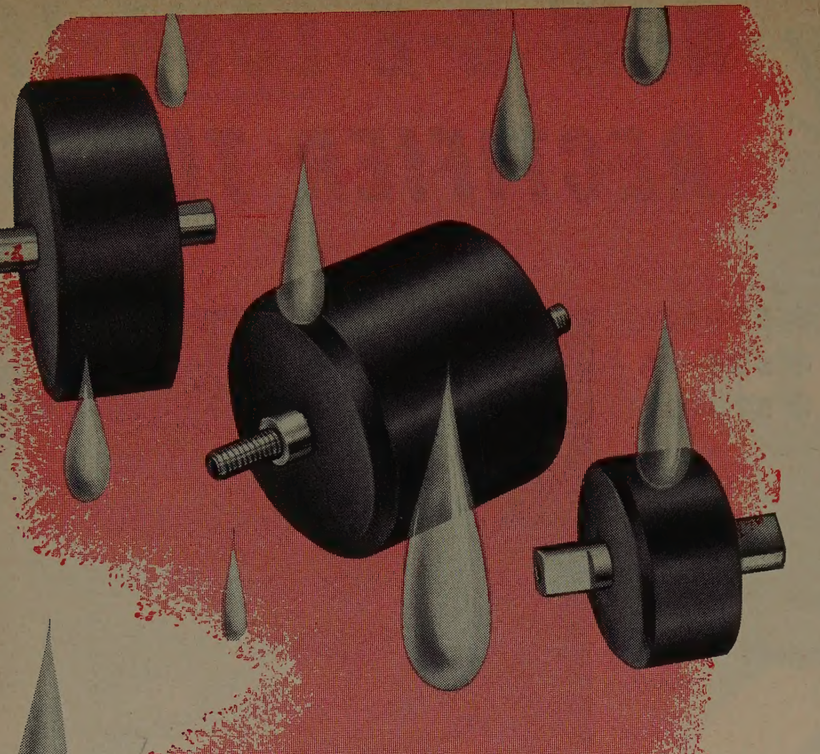
Base construction now is button-stem, with the leads passing through widely spaced individual seals at bottom of glass envelope. Adds strength, gives shorter leads and greater lead separation, and brings about better heat conduction . . . which in turn reduces electrolysis and air-leakage.

New G-E 5U4-GA has same base diameter and layout as prototype 5U4-G—is fully interchangeable.

Progress Is Our Most Important Product

GENERAL ELECTRIC

162-1A5



All "SLUGS" look alike, but . . .

Hi-Q® CARTWHEELS*

have that unique sealing!

FUNCTION FITTED

Heavy ceramic body; positively-bonded electrodes; intimately-joined terminals—such details are common to all "slug" ceramic capacitors. The assembly is then sealed—and that's where Hi-Q "Cartwheels" are **different.**

"Cartwheels" feature a **cast casing,** completely and permanently sealed in one operation. The exclusive potting compound results in meticulous jacketing.

Especially developed for Color-TV, Hi-Q "Cartwheels" mean ratings up to 30 KV; much higher corona-starting voltages; greatly increased dielectric strength; excellent arc-resistant properties; insulation resistance greater than 50,000 megohms; power factor of 1.5% max. at 1000 cps; greatest immunity to humidity and heat; outstanding service life.

Get the FACTS!

Latest literature on request. Hi-Q specialists will gladly collaborate on your high-voltage and other ceramic-capacitor requirements.

*Reg. trade-mark

Hi-Q®
DIVISION

AEROVOX CORPORATION
OLEAN, N. Y.

AEROVOX CORPORATION
NEW BEDFORD, MASS.

ACME ELECTRONICS, INC.
MONROVIA, CALIF.

CINEMA ENGINEERING CO.
BURBANK, CALIF.

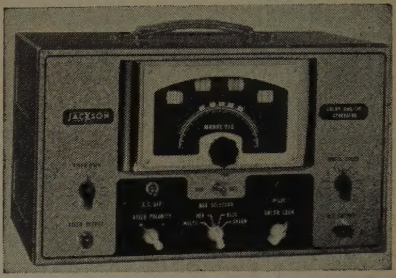
In Canada: AEROVOX CANADA LTD., Hamilton, Ont.
JOBBER ADDRESS: 740 Belleville Ave., New Bedford, Mass.

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 27A)

nounced by the **Jackson Electrical Instr. Co.,** 16 S. Patterson Blvd., Dayton 2, Ohio. Designated the Model 712, the new generator is priced at approximately **$350.**



The instrument is entirely self-contained and provides a complete NTSC system color difference signal as well as all required synchronizing signals. A 4.5 mc crystal controlled oscillator provides for accurate tuning of the color set's fine tuning control. A 3.58 mc crystal controlled burst oscillator, with its associated Color Lock control, assures accurate locking with the color burst generator in the set.

When the generator is set to produce Color Bars, either single bars or a multibar pattern are available. Single bars, selected from a front panel switch include Red, Blue, or Green bars. In the "Multi" position the instrument produces 5 simultaneous bars consisting of Orange (I signal), Red (R-Y), White (multiple), Magenta (Q) signal, and Blue (B-Y).

The new Jackson generator will also produce a white dot pattern made up of 6 rows of 8 dots each for the convergence adjustments of tri-color tubes. A cross-hatch pattern is also available for linearity adjustments.

Laminations Catalog

Magnetics, Inc., Butler, Pa., has just issued a new Magnetic Lamination Catalog describing the company's standard line of laminations, laminated cores, and discs.

Catalog ML 101, "Performance Guaranteed Magnetic Laminations," includes lamination specification sheets, showing both the individual laminations to actual scale, as well as properties of square cross-section core stacks, and weights and counts for different materials.

Sections are devoted to laminated core assemblies, mechanical and magnetic parameters and lamination tolerances.

The firm manufactures both progressive tungsten-carbide dies for high production, and fine tolerance laminations and hi-carbon, hi-chrome dies for shorter production runs. These dies for "Performance Guaranteed" laminations are described. The catalog also includes a section on the proper information needed in determining and ordering quantities of magnetic laminations.

Copies may be obtained by writing on company letterhead.

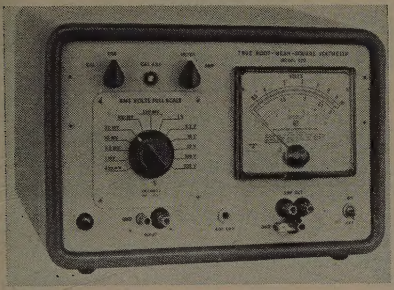
News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 32A)

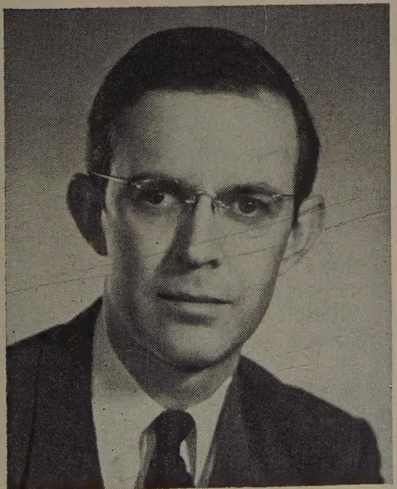
RMS Voltmeter

Model 230, manufactured by **Ballantine Labs., Inc.,** Boonton, N. J., is a voltmeter of the ultra-sensitive type, designed to measure true root-mean-square values of highly complex wave forms, and sinusoidal waves. It operates over a range of 100 mv to 320 volts and in a band of 5 cps to 500 kc. Accuracy is better than 3 per cent between 15 cps and 150 kc for any reading regardless of scale position.



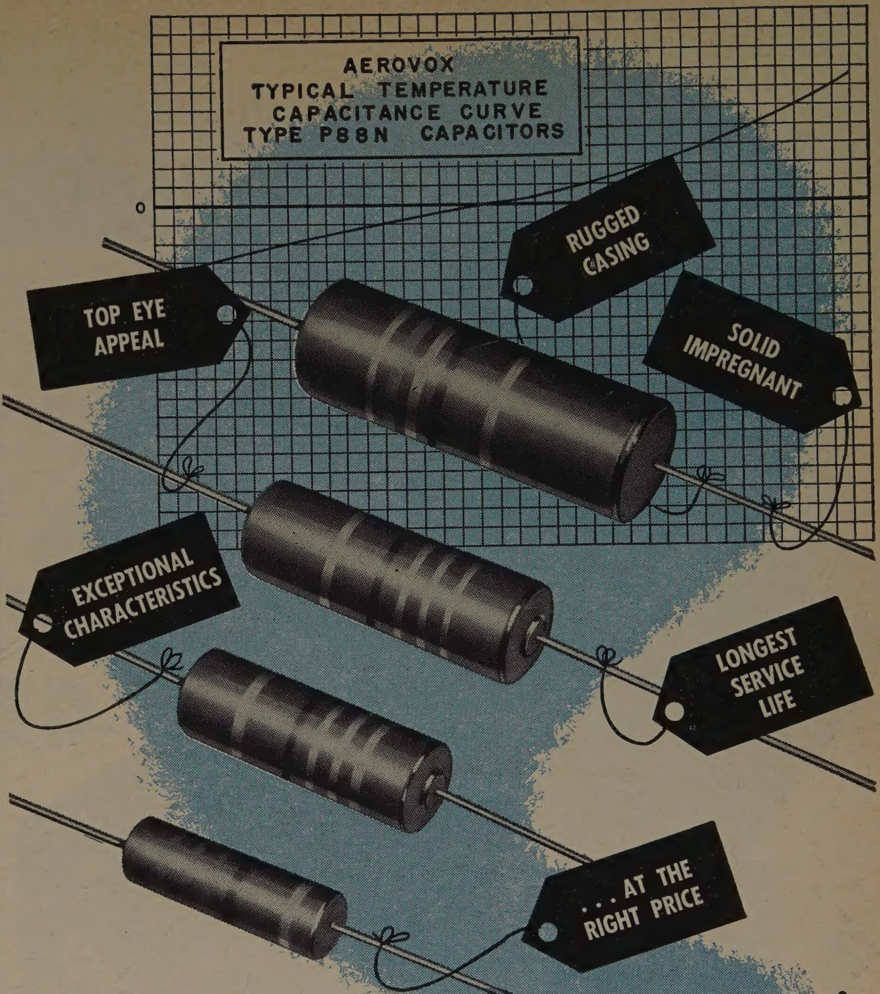
Other features include 10 megohm input impedance, provision for simultaneously observing the voltage reading and monitoring the amplified signal with phones or CRO, and a built-in calibrator unit for correcting the effects of advanced aging of tubes. Accessories are available for extending voltage range to 20 mv and 10 kv and for measuring rms currents from 0.1 microampere to 10 amperes.

Boyers Heads National Recorder Division



The appointment of Mr. John S. Boyers to the **National Co.,** Malden, Mass., as Chief Engineer of the Magnetic Memory Devices Division was recently announced by Paul H. Frye, Vice-President in charge of Engineering.

(Continued on page 74A)



DURANITE*

MOLDED TUBULAR PAPER CAPACITORS

The **new** and **improved** Duranite (Type P88N) paper tubulars still feature **Aerolene***—the **solid** impregnant—combining the advantages of wax and oil impregnants. No need of stocking both types. No risk of shelf deterioration.

And now Duranites are molded in **blue** non-inflammable plastic. Top eye appeal—and outstandingly rugged. Pigtails, centered and firmly imbedded, won't work loose or pull out.

Units essentially immune to moisture penetration. Exceptional performance characteristics—insulation resistance; power factor vs. temperature; 100° C. operating temperature.

Get the FACTS!

Descriptive literature on request. Write on business letterhead for sample. Standard values stocked for immediate delivery. Let us quote on your needs.

FUNCTION FITTED

AEROVOX capacitors

AEROVOX CORPORATION

NEW BEDFORD, MASS.

Hi-Q DIVISION OLEAN, N. Y.	ACME ELECTRONICS, INC. MONROVIA, CALIF.	CINEMA ENGINEERING CO. BURBANK, CALIF.

Export: Ad Auriema, Inc., 89 Broad St., New York, N. Y. Cable: Auriema, N. Y.

*Trade-mark

Heiland RECORDING **700C Multi-Channel OSCILLOGRAPH**

The Key to Better Product Design-Performance

Heiland's new 700C Recording Oscillograph enables the testing engineer and scientist to solve a wide variety of industrial and laboratory problems involving the measurement and correlation of strains, stresses, vibrations, accelerations, pressures, impacts, temperatures, etc.

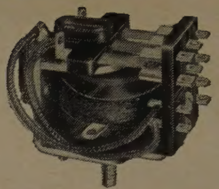
The 700C Oscillograph provides up to 60 separate recording channels, has record widths of 8 to 12 inches, record speeds as high as 144 inches per second and as low as .030 inches per second. The 700C accommodates Heiland's new Sub-Miniature Galvanometers and temperature controlled magnet assemblies in which a new high in stability and sensitivity has been attained.

Heiland
20TH ANNIVERSARY

Heiland Research Corporation 130 EAST FIFTH AVE.
DENVER, COLORADO

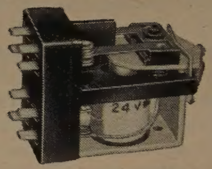
high quality at low cost!

3 NEW
Potter & Brumfield
RELAYS



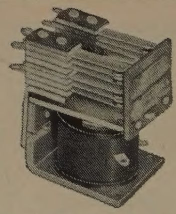
GENERAL PURPOSE
SERIES KA

Designed for either current or voltage actuation. Small size, high capacity and many contact combinations make KA series universal in application.



APPLIANCE SERIES AB

The AB relay is primarily designed for appliance applications to eliminate noisy and troublesome clapper or solenoid type contactors.



MULTIPLE LEAF
SERIES GA

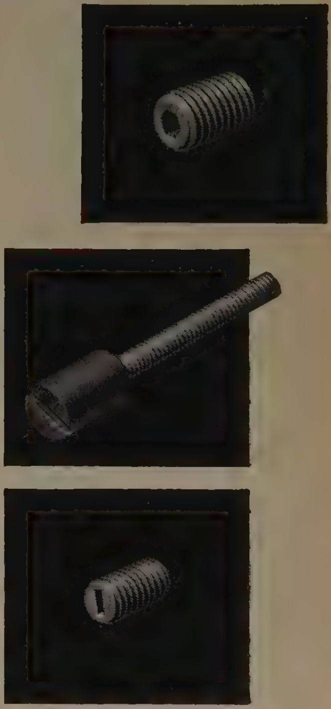
The GA series permits a wide range of contact combinations for multiple circuit switching of power loads.

samples available for immediate shipment • quotations on request

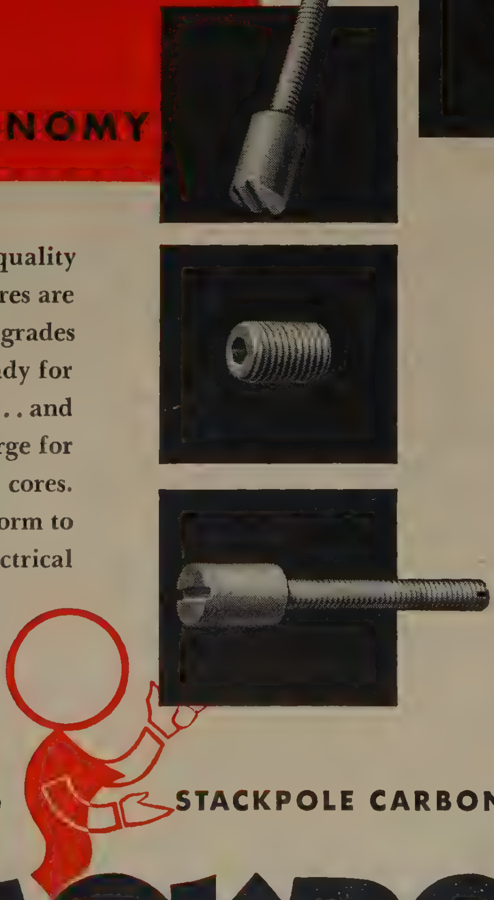
POTTER & BRUMFIELD
PRINCETON, INDIANA
subsidiary of AMF
NEW YORK, N. Y.
Export: 13 E. 40th St., New York
Sales offices in principal cities

CUT IRON CORE COSTS

with Stackpole "PREFERRED TYPES"



"EE" SERIES . . . FOR ENGINEERED ECONOMY



Made to well-known Stackpole quality standards, these new "EE" Cores are available only in commonly needed grades and sizes. They're ready for delivery from stock . . . at low prices . . . and without the usual set-up charge for custom-engineered cores. Mechanical specifications conform to the latest MPA recommendations. Electrical standards fully meet 8 out of 10 requirements of radio, TV, and communications equipment. Write, wire, or 'phone for details.

Electronic Components Division

STACKPOLE CARBON COMPANY
St. Marys, Pa.

STACKPOLE

BLILEY
BX
33/64"

. . . Masterpieces in crystal craftsmanship

Bliley
BANTAM BX
CRYSTAL UNIT

BANTAM IN EVERY WAY—
BUT PERFORMANCE!

The **BANTAM BX** is precision in pint-size! Meets all of the demanding specifications dictated by space limitations in portable and pocket size gear.

In design, if the sub-miniature assembly calls for multi-channel operation, **BANTAM BX** crystals with wire leads may be conveniently mounted in a sub-miniature selector switch. Or, the **BANTAM BX** can be directly wired into printed circuitry.

BANTAM BX crystals meet all performance requirements of larger units such as the Bliley BH6A. Supplied in frequency ranges: 10 mc - 100 mc. Hermetically sealed.

More technical information may be secured by requesting our Bulletin No. 46-A. Send your prints for a prompt quotation.

TECHNICAL LITERATURE AVAILABLE:
Bulletin No. 47—Crystals, Ovens, Frequency Standards, MIL-type Specification Index
Bulletin No. 45-A—Solid Ultrasonic Delay Lines
Bulletin No. 46-A—"Bantam BX" Crystals
Bulletin No. 44-B—Amateur, Standard Frequency, Ship-To-Shore, and TV Service Crystals

Bliley
CRYSTALS

Bliley
ELECTRIC COMPANY
UNION STATION BUILDING
ERIE, PENNSYLVANIA

Industrial Engineering Notes

AERONAUTICS*

Plans were completed by the Civil Aeronautics Administration and the Air Force for CAA operation of Radar Approach Control Centers (RAPCONS) at 18 military bases, to serve both civil and military traffic, it was announced by F. B. Lee, CAA Administrator. Under the program, the CAA will operate the surveillance radar units for approach control of all traffic in the vicinity, while the Air Force will operate the precision approach radar for actual landings by military craft at the air bases under instrument flight rules (IFR) conditions. All equipment for the RAPCONS will be provided by the Air Force, which consists of three surveillance and two precision approach radar scopes. The CAA also announced the procurement of 75 pieces of equipment for two-way recording at some 25 busy towers and centers around the country. The equipment has already been installed at the Washington National Airport, now making it possible for the first time to record all conversations between pilots and CAA tower and center personnel. Previously, only outgoing CAA messages were recorded.

FCC ACTIONS

The Federal Communications Commission adopted, with modifications, proposals made September 4, 1952, to provide for the operation of certain fixed radio (nonbroadcast) stations in the 72-76 mc band and, at the same time, afford adequate protection to television operations on Channel 4 (66–72 mc) and Channel 5 (76–82 mc) which straddle that band. The amended rules were effective August 9, 1954. The resulting standards, designed to minimize the possibilities of interference to television channels 4 and 5, provide (1) that all operations in the 72–76 mc band are subject to the condition that no harmful interference will be caused to TV reception on the channels; (2) present fixed operations in that band will be permitted to continue on a noninterference basis, but interference complaints must be cleared up within 90 days; (3) that a fixed station would be authorized between 10 and 80 miles from such a transmitter only if it came within certain criteria established by charts provided for this purpose; but (4) that no fixed station would be authorized for operation in this band when the proposed separation between its transmitter and that of a TV station transmitter operation on Channel 4 or 5 was 10 miles or less.

FEDERAL PERSONNEL

Several high-ranking Signal Corps officers have been reassigned, Maj. Gen. George I. Back, Chief Signal Officer, announced. Brig. Gen Wesley T. Guest,

(Continued on page 48A)

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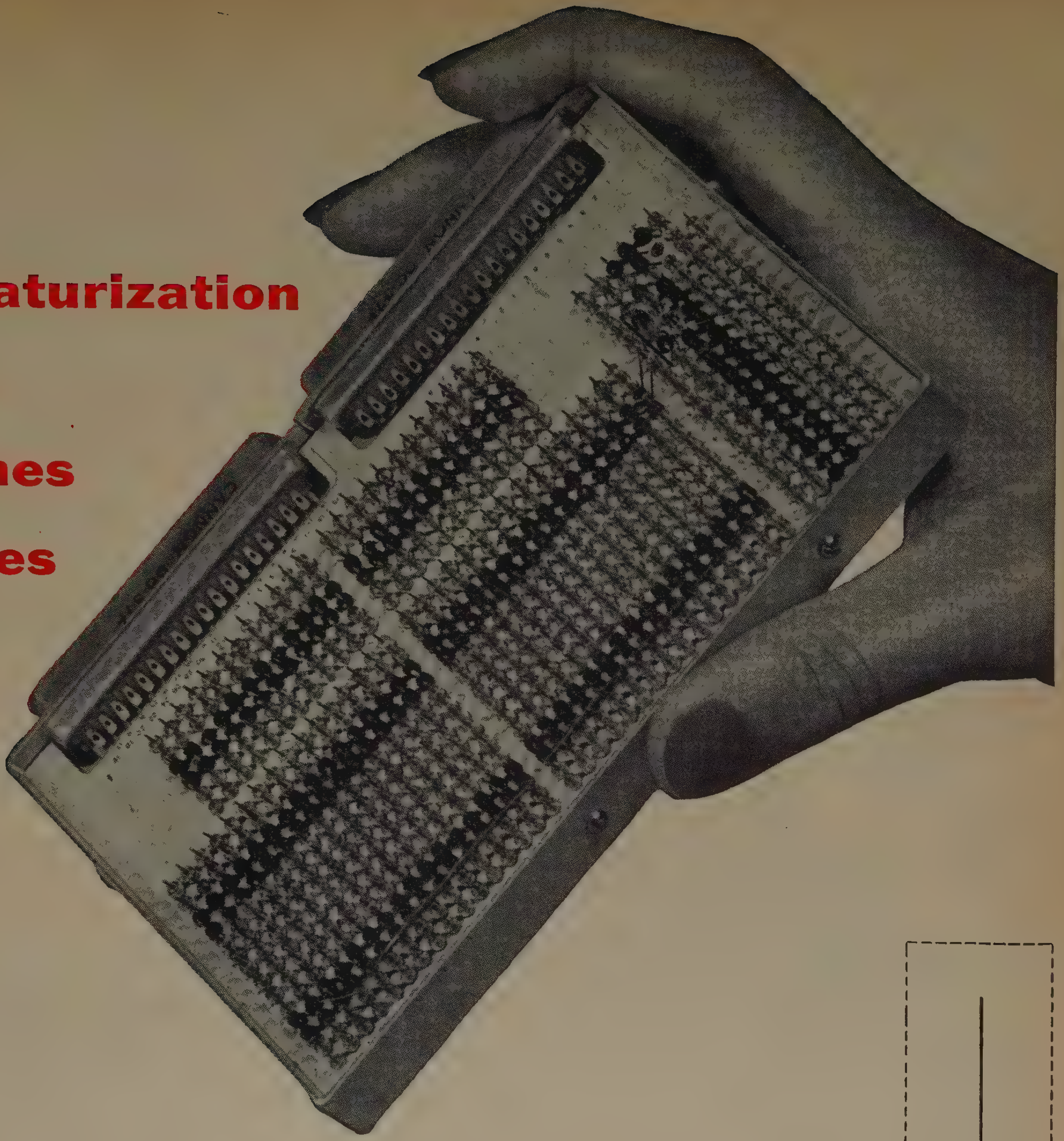
* The data on which these NOTES are based were selected by permission from *Industry Reports*, issues of July 5, 12, 19, 26, and August 2, published by the Radio-Electronics-Television Manufacturers Association, whose helpfulness is gratefully acknowledged.

Miniaturization with Hughes Diodes



New computer matrix has high component density

This experimental reading gate matrix for airborne computers effectively *utilizes the subminiature size of Hughes Point-Contact Germanium Diodes*. Developed by the Miniaturization Group of Hughes Research and Development Laboratories, the unit measures 5¼ by 3⅛ by ½ inches (excluding plugs and frame). It contains 504 diodes, 209 resistors.* Average component density: *94.5 per cubic inch!*

Frequently, space requirements of conventional wiring techniques will not permit electronic equipment to be miniaturized to the same extent as the components. However, spot-welded connections can effectively reduce wiring space . . . and it is easy to spot-weld the dumet leads of Hughes diodes. There is no adverse effect on diode characteristics, even when the connections are welded close to the diode body. With Hughes diodes, designers can take full advantage of advanced packaging and wiring techniques.

Hughes diodes are easy to mount in conventional assemblies or in subminiature equipment. In service, these diodes have earned a reputation for reliable performance and stability under severe operating conditions. Make your selection from the many standard and special types available — all listed and described in our new Bulletin, SP-2A.

*Actual size, diode body: 0.265 by 0.130 inches, maximum.

Reprints of a paper describing the packaging techniques of the subminiature matrix are available, too. Your copy will be sent promptly on request.

Hughes
SEMICONDUCTOR SALES DEPARTMENT
Aircraft Company, Culver City, Calif.
New York Chicago

NO GUESSWORK

7 3 4

NO CHANCE FOR ERROR

with a
DIRECT READING
HYCON
DIGITAL
VTVM
MODEL 615

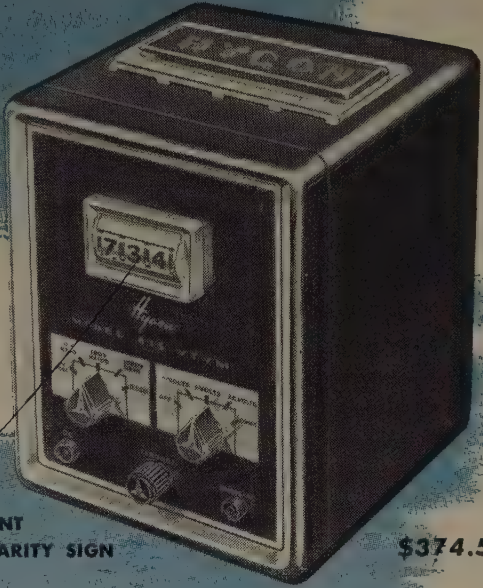
ILLUMINATED DECIMAL POINT
AND POLARITY SIGN

$374.50

The Model 615 VTVM is a precision instrument — functional in design . . . professional in appearance. The direct-reading digital display eliminates most interpolation error — shortens costly "learning curve" in factory and assembly line inspection. Other features — never before offered in an instrument of comparable price — include 1% accuracy (DC and ohms), and *1 millivolt* sensitivity. Inspect the Model 615 at your Electronic Parts Jobber's. You'll agree the new standard is Hycon . . . "where accuracy counts."

• 12 RANGES: AC, DC, OHMS • AC FREQUENCY RESPONSE TO 250 MC (with auxiliary probes) • OVERLOAD PROTECTION • LIGHTWEIGHT, STURDY STEEL CASE • PROVISIONS FOR BENCH STACKING

The Model 615 VTVM is one of a matching set of precision test instruments, which includes the Model 617 Oscilloscope (designed for color TV) and the Model 614 Standard VTVM.



Service facilities in your area.

Hycon Mfg. Company
2961 EAST COLORADO STREET, PASADENA 8, CALIFORNIA
"Where Accuracy Counts"

Industrial Engineering Notes

(Continued from page 36A)

formerly Commandant of the Army Signal School at Fort Monmouth, N. J., will become Chief Signal Officer of the European Command, succeeding Brig. Gen. Edward R. Petzing, who is returning to the Office of the Chief Signal Officer.

Brig. Gen. Victor A. Conrad, Chief of the Procurement and Distribution Branch, OCSO, will succeed Gen. Guest in the Fort Monmouth School. Also returning to the Office of the Chief Signal Officer is Brig. Gen. Elton F. Hammond, who has been Signal Officer for Army Forces, Far East. He will be replaced in the Far East assignment by Col. Albert F. Cassevant, who was Assistant Chief of the Procurement and Distribution Division of the OCSO. Col. William D. Hamlin, Army representative on the staff of the Defense Department's Office of Transportation and Communications, has been named Assistant Commander of the Signal Corps School at Fort Monmouth. Succeeding Col. Hamlin will be Col. Kenneth F. Zitzman. Capt. John T. Hayward (USN) has been named Commanding Officer of the Naval Ordnance Lab., White Oak, Md. He succeeds Capt. Edward L. Woodyard, who retired after 30 years of service.

INDUSTRY STATISTICS

Television production during the first half of this year was reported as 2,845,147 units compared with 3,834,236 sets manufactured in the same 1953 period. The six-month radio production total was 4,886,559 compared with 7,265,542 sets manufactured in the comparable 1953 period. The report showed the production of 8,394 color sets in the first six months, of which 347 were manufactured in June. The number of sets manufactured this year to tune uhf increased to 636,456 with the production of 99,404 receivers in June.

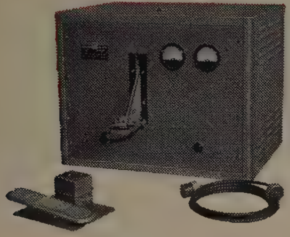
MOBILIZATION

The Bureau of Aeronautics, Department of the Navy, announced a new policy in which its various divisions having cognizance over design and procurement of electronics are directed to encourage contractors to produce equipment utilizing mechanized production techniques. This is particularly true, in the procurement of equipments with high mobilization requirements. "Mechanically produced electronic equipment has shown many superior military qualities compared to equipment produced by conventional means," according to the BuAer announcement, "and progress in the field of mechanical production has now advanced sufficiently to warrant its use in field equipment for the fleet." The newly-announced policy, it was pointed out, "provides substantial incentive for manufacturers to use the BuAer-Bureau of Standards developed 'Tinkertoy' process."

Continued on page 58A)

marion

advancement in instrument design

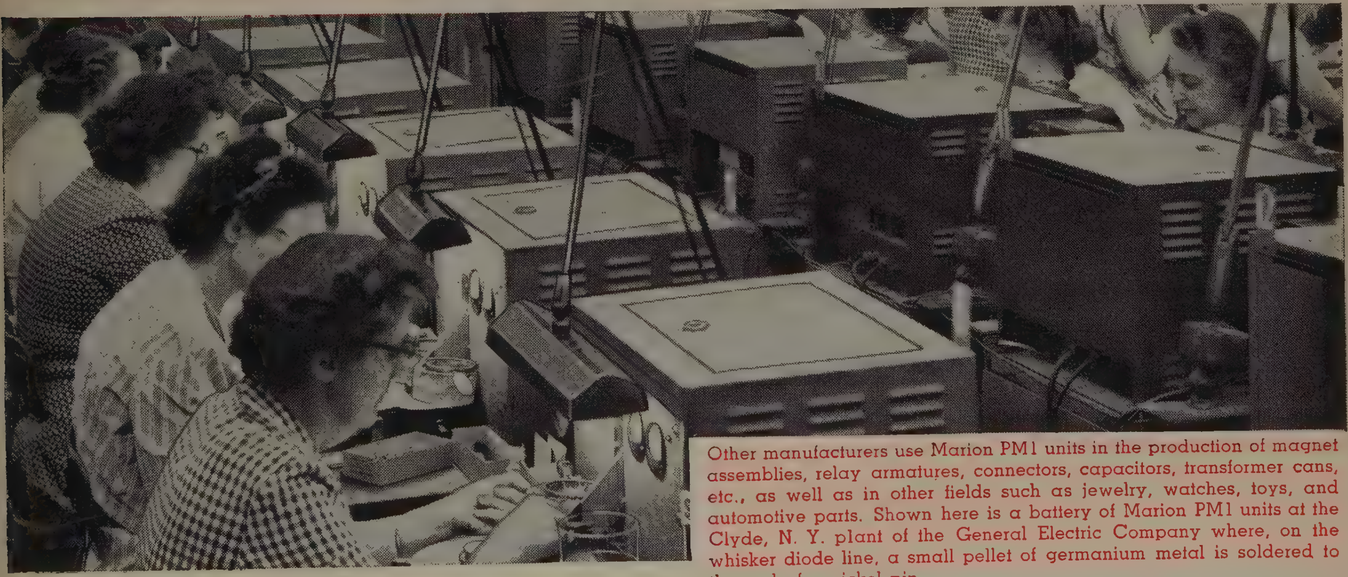


MARION MODEL PM1 INDUCTION SOLDERING UNIT

Low cost, low powered and portable. . . . Size 15¾″ x 21½″ x 15″. 150 lbs. Power supply: 115 volts 60 cycles. Draws 775 watts full load, 100 watts standby.

IMPROVED SOLDERING OF SMALL PARTS AND ASSEMBLIES

Typical of the Marion developments that have helped make Marion stand for "advancement in instrument design" is the Marion Model PM1 Induction Soldering Unit. Originally designed and presently used by Marion for true glass-to-metal hermetic sealing of Marion meters, it also has proven to be a valuable production tool for many purposes. Illustrated above, for example, is Marion's use of the PM1 in the soldering of magnet assemblies where quality and uniformity result.



Other manufacturers use Marion PM1 units in the production of magnet assemblies, relay armatures, connectors, capacitors, transformer cans, etc., as well as in other fields such as jewelry, watches, toys, and automotive parts. Shown here is a battery of Marion PM1 units at the Clyde, N. Y. plant of the General Electric Company where, on the whisker diode line, a small pellet of germanium metal is soldered to the end of a nickel pin.

MILLIAMPERES

The Marion Model PM1 speeds up production, reduces costs and improves quality. Heat is generated within the work itself — even in parts otherwise inaccessible. Oxidation, scaling and damage to surface finish are minimized. Soldering of an entire seam or several jig-located parts at one time is readily accomplished.

This is an example of how Marion's belief in "Advancement in Instrument Design" has produced a production tool which not only improves Marion instruments but also provides other manufacturers with better soldering equipment. **Marion Electrical Instrument Company, 407 Canal Street, Manchester, N. H.**

marion meters

Reg. U. S. Pat. Off.

MANUFACTURERS OF RUGGEDIZED AND "REGULAR" METERS AND RELATED PRODUCTS

Copyright 1953 Marion Elec. Instr. Co.

Ruggedized
and aged

"RELIABLE" DOUBLE TRIODE

Do you have an aircraft or industrial application that requires *utmost* dependability in increasing or controlling alternating voltages or powers . . . in changing electrical energy from one frequency to another . . . or in generating an alternating voltage?

If so, specify the Red Bank RETMA 6385 "Reliable" Double Triode. For it is specially ruggedized to perform at top efficiency longer, even under operating conditions of severe shock and vibration. And, as further assurance of its extra reliability, each RETMA 6385 is factory-aged with a 45-hour run-in under various overload, vibration and shock conditions, such as it might meet on the job.

Whether you need tubes as amplifiers, mixers, or oscillators, it will pay you to investigate the superior, longer-lasting performance qualities of the Bendix Red Bank RETMA 6385.

RATINGS*

Heater voltage—(AC or DC)**	6.3 volts
Heater current	0.50 amps.
Plate voltage—(max.)	360 volts
Max. peak plate current (per plate)	25 ma.
Max. plate dissipation (per plate)	1.5 watts
Max. peak grid voltage	+ 0 volts −100 volts
Max. heater-cathode voltage	300 volts
Max. grid resistance	1.0 megohm
Warm-up time	45 sec.

(Plate and heater voltage may be applied simultaneously.)

*To obtain greatest life expectancy from tube, avoid designs where the tube is subject to all maximum ratings simultaneously.

**Voltage should not fluctuate more than ±5%.

PHYSICAL CHARACTERISTICS

Base	Miniature button 9-pin
Bulb	T-6½
Max. over-all length	2 3/16 in.
Max. seated height	1 15/16 in.
Max. diameter	7/8 in.
Mounting position	Any
Max. bulb temp.	160° C

AVERAGE
ELECTRICAL CHARACTERISTICS

Heater voltage, E_f	6.3 volts
Heater current, I_f	0.50 amps.
Plate voltage, E_b	150 volts
Grid voltage, E_c	−2.0 volts
Plate current, I_b	8.0 ma.
Mutual conductance, g_m	5000 μmhos
Amplification factor, μ	35
Cut-off voltage	−10 volts
Direct interelectrode capacitances (no shield)	
Plate-grid (per section)	1.7 $\mu\mu$f
Plate-cathode (per section)	1.1 $\mu\mu$f
Grid-cathode (per section)	2.4 $\mu\mu$f
Plate-plate	0.1 $\mu\mu$f

Bendix Red Bank
ECLIPSE

Manufacturers of Special-Purpose Electron Tubes, Inverters, Dynamotors and Fractional HP D.C. Motors

DIVISION OF

THE NAME Bendix AVIATION CORPORATION MILLIONS TRUST

EATONTOWN, N. J.

West Coast Sales and Service: 117 E. Providencia, Burbank, Calif.
Export Sales: Bendix International Division, 205 E. 42nd St., New York 17, N. Y.
Canadian Distributor: Aviation Electric Ltd., P.O. Box 6102, Montreal, P.Q.

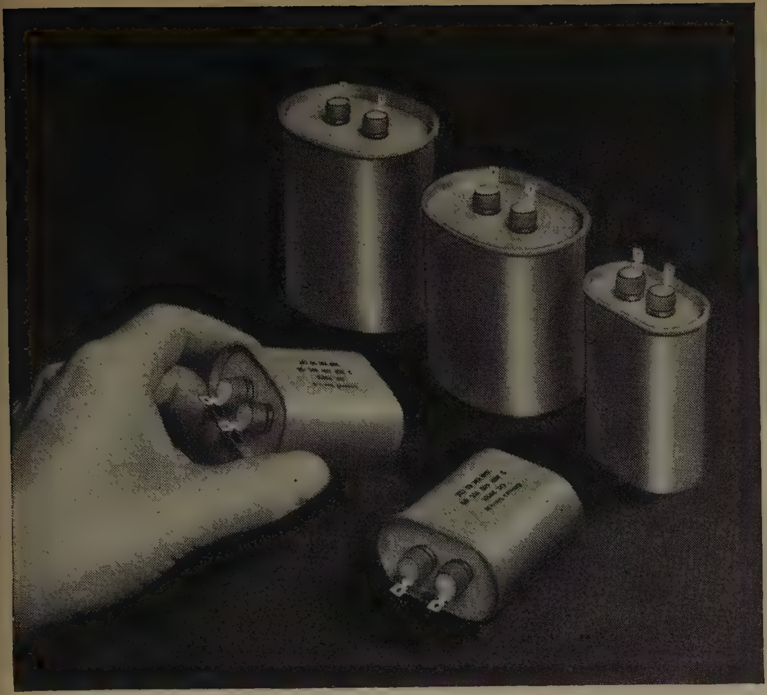
Industrial Engineering Notes

(Continued from page 48A)

RETMA Activities

The Joint Technical Advisory Committee, under the chairmanship of A. V. Loughren, Hazeltine Corp., has filed a 10-page statement with the Federal Communications Commission regarding the Commission's proposed rule-making proceedings on restricted radiation devices. JTAC, a technical committee formed by RETMA and the IRE to advise the government, commented on the FCC docket (9288) and brought the Commission up-to-date on the work done by the committee on spurious emissions in connection with an earlier government request. On the general subject, JTAC advised the FCC that it is continuing a study program to determine presently and prespectively: (1) the intensity of interfering signal that each service in each portion of the spectrum could feasibly be expected to tolerate; (2) restrictions on spurious radiations which would be necessary to avoid interference of more than these maximum amounts; (3) technical factors governing the cost of alternative methods of meeting the resulting requirements; (4) situations which cannot feasibly be met by initial design or type modifications of apparatus. On completion of this program, JTAC told the FCC it would then be in a position to propose limitation standards for spurious radiations by categories. Commenting on the specific proposals in the FCC rule-making proceedings, JTAC felt that the Commission's definition of "harmful interference" was too broad. JTAC also said that it felt that the FCC suggestions relating to field strengths permissible after June 30, 1956, were not based on known technical data. It also was pointed out that the committee did not feel that this reduction in field strength would solve an appreciable number of interference cases. JTAC suggested that the FCC might wish to consider withholding adoption of its Table until after June 30, 1956, and to state that periodic review will be made taking into consideration all new developments in the art. The technical committee noted that in some areas standardized measurement methods for spurious radio emission are not yet available. In conclusion, JTAC said it will continue its investigations and that it hoped that before the June 1956 effective date definite information permitting a more factual approach to these problems will be available. . . . **A Symposium on Printed Wiring and Circuitry, under sponsorship of the RETMA Engineering Department, has been scheduled on January 27–28, 1955, at the University of Pennsylvania, Philadelphia, Pa.** A task group from the Engineering Department met on August 4 in the New York RETMA offices to make plans for the forthcoming symposium. . . . **RETMA went on record as opposing a suggestion that the Attorney General be approached by a congressional subcommittee for an**

(Continued on page 60A)



RATINGS of G-E drawn-oval capacitors range from 1 to 10 uf, 600 to 1500 volts d-c, and 330 to 660 volts a-c.

G-E drawn-oval capacitors save space and cost less

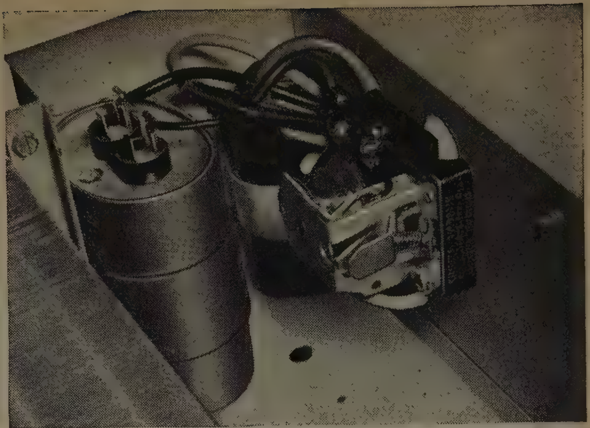
If you use fixed paper dielectric capacitors, G-E drawn-ovals offer you an opportunity to save up to 20% on weight, and as much as 10% to 20% on cost. The oval-shaped container, developed by General Electric, offers more capacitance per dollar than similarly rated rectangular capacitors. And, by conforming to the natural shape of the winding, it results in a smaller, lighter unit, too. They're available in ratings from 1 to 10 uf, 600 to 1500 volts d-c, or 330 to 660 volts a-c, 60 cycles.

G-E drawn-ovals feature: A double-rolled seam, between case and cover that makes a mechanically strong, hermetic seal which stays leak-proof even under severe operating conditions; a choice of eyelet, fork-type, or quick-connect (solderless) terminals; silicone bushings between terminal and cover, that effectively maintain a high insulation resistance despite long operation and wide temperature variation.

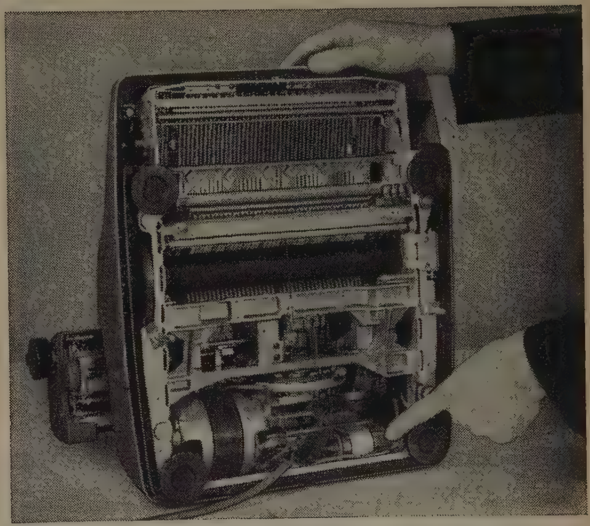
General Electric drawn-oval capacitors are being used in room air conditioners, business machines, fluorescent lighting ballasts, and industrial and military control systems. If you would like specific application assistance, contact your local General Electric Apparatus Sales Office. General Electric Company, Schenectady 5, New York.

Progress Is Our Most Important Product

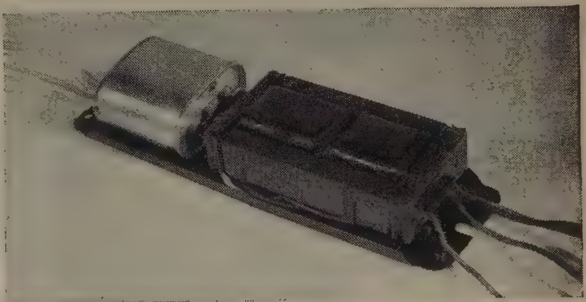
GENERAL ELECTRIC



IN ROOM AIR-CONDITIONERS G-E drawn-oval improves power factor and reduces running current.



IN ELECTRIC TYPEWRITERS AND BUSINESS MACHINES, the compact G-E drawn-oval is used with split-phase capacitor-run motors.



IN FLUORESCENT LAMP BALLASTS, G-E drawn-ovals (left) improve power factor.

CAN YOU USE THIS VERSATILE CAPACITOR?

General Electric Co.
Section D442-19
Schenectady 5, New York

Please send me Bulletin GEA-5777.

Name____________________

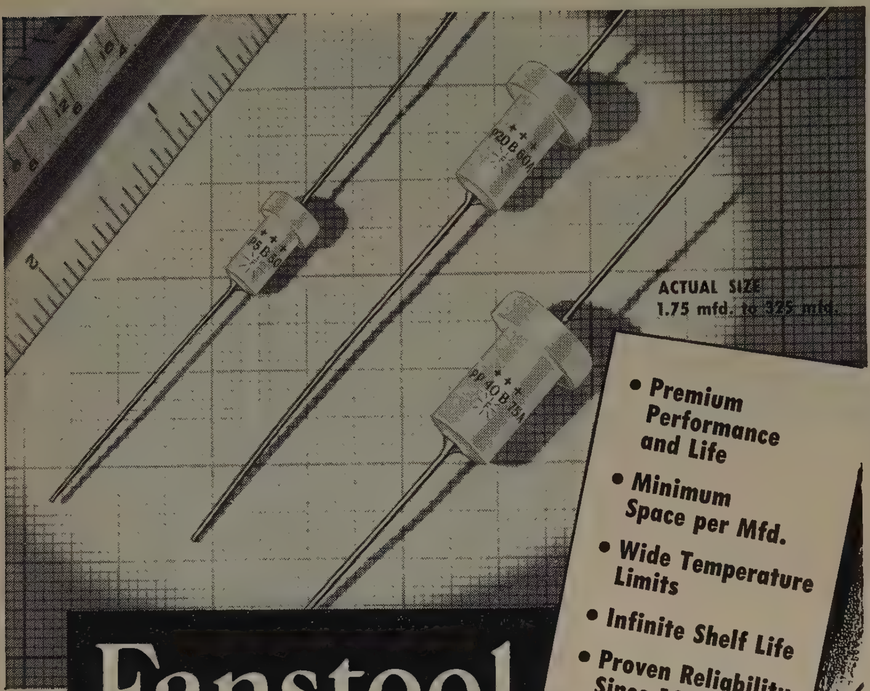
Position____________________

Company____________________

City____________ Zone___ State________

TANTALUM CAPACITORS...

...basic in current electronic trend..



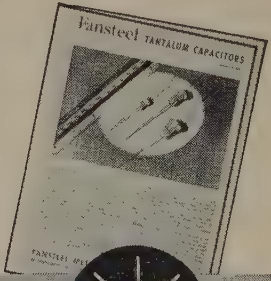
ACTUAL SIZE
1.75 mfd. to 325 mfd.

- Premium Performance and Life
- Minimum Space per Mfd.
- Wide Temperature Limits
- Infinite Shelf Life
- Proven Reliability Since 1930

Fansteel
TANTALUM
CAPACITORS

Now, through the use of tantalum, new high standards of electrolytic capacitor performance are available. The tantalum oxide film is the most stable dielectric, chemically and electrically, yet discovered. As a result, Tantalum Capacitors offer advantages not found in any other electrolytic type — long life, space saving, wide temperature range excellent frequency characteristics, no shelf aging.

Tantalum Capacitors are made by Fansteel and other leading capacitor manufacturers. Ask for current information bulletins on Fansteel Tantalum Capacitors.



FANSTEEL METALLURGICAL CORPORATION
NORTH CHICAGO, ILLINOIS, U. S. A.
Tantalum Capacitors... Dependable Since 1930

32503C

Industrial Engineering Notes

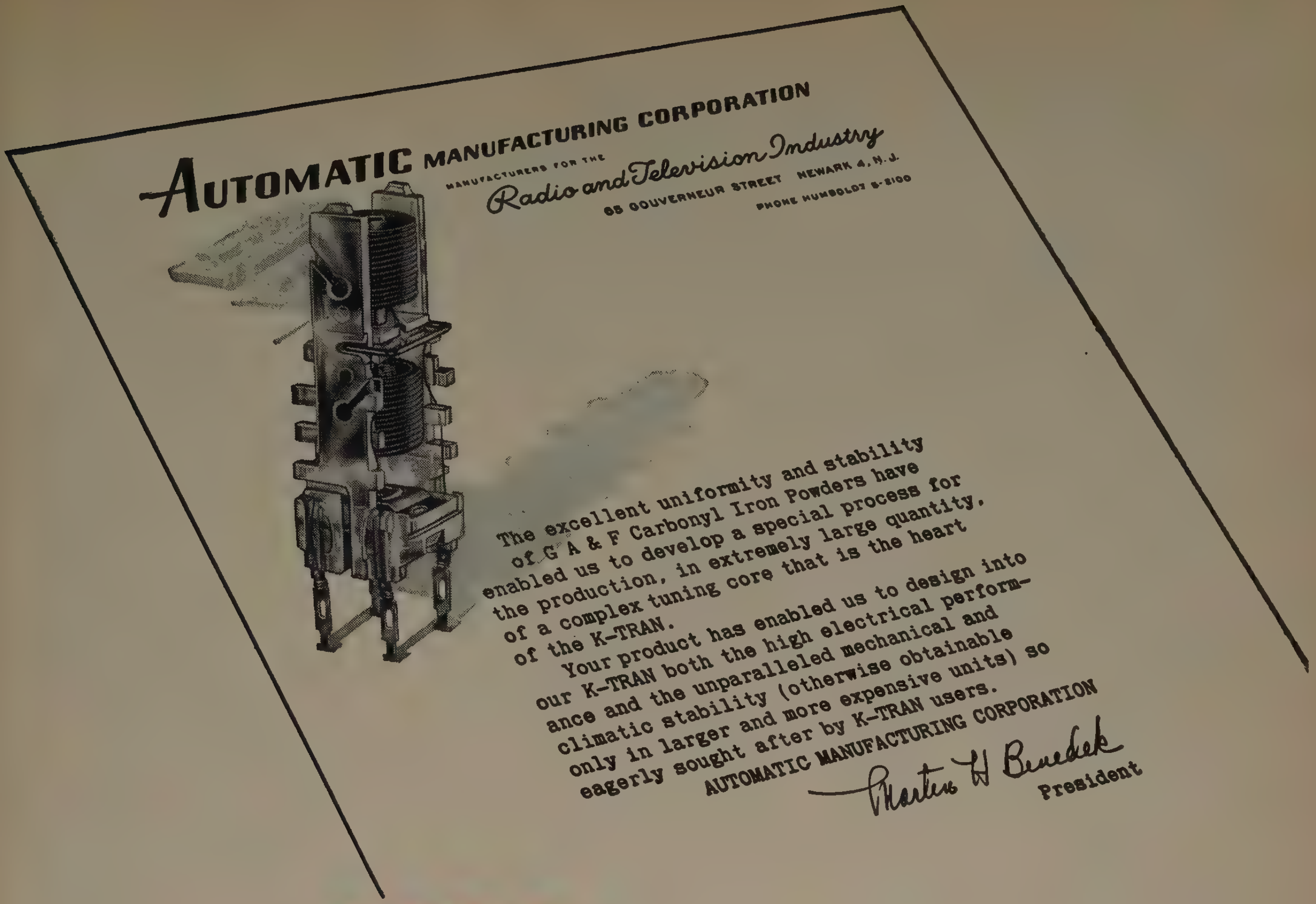
(Continued from page 58A)

opinion looking toward antitrust exemption for television set manufacturers who agree to produce only all-channel receivers. The Association also expressed its opposition to a suggestion which would ban the interstate shipment of TV sets which are not capable of all-channel reception. The comments were submitted in a letter by RETMA President Glen McDaniel to Senator Charles Potter (R., Mich.), chairman of the Subcommittee on Communications of the Senate Interstate and Foreign Commerce Committee. The subcommittee has conducted hearings recently on the status of uhf telecasting and the suggested measures were presented by two different witnesses who appeared before the group. Mr. McDaniel said he would consider it a "serious mistake" for the subcommittee to seek an opinion from the Attorney General looking toward an antitrust exemption for set manufacturers who agree to make only all-channel receivers. He pointed out in his letter to Senator Potter that: "The basic purpose of the antitrust laws is to preserve free competition. The members of this industry are in full sympathy with that objective and believe that the antitrust laws are a necessary and valuable part of the laws of the United States. This industry, which has been referred to with good reason as being 'the most competitive of American industries,' adheres to both the spirit and the letter of the federal antitrust laws. As I stated in my previous testimony (before the subcommittee): 'We believe that a system of free competition is best for the public in the long run.' In short, we believe any weakening of the application of the antitrust laws is contrary to the best interests of the American public." As to the suggestion that the powers of Congress be used to prohibit the interstate shipment of TV sets which are not equipped for all-channel reception, Mr. McDaniel said: "We believe that acceptance of the suggestion and exercise of such a power by Congress would be an abuse of federal power, would mean the intervention by government in the regulation of the manufacture of television sets and would be of doubtful constitutionality. Each of these consequences in itself is ample justification for rejection of the suggestion." The set manufacturers are aware of the problems confronting many broadcasters today, Mr. McDaniel said, and are vitally concerned with their well being. He told Senator Potter that although it is the hope of the industry that the broadcasters' problems can be solved: "Under no circumstances do we believe that it would be proper or desirable to solve the problems by carving out exceptions to the antitrust laws or by imposing federal regulation on the manufacturer of television sets."

TECHNICAL

The National Bureau of Standards issued a new publication covering the second phase of a continuing program for the de-

(Continued on page 68A)



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Charles H Benedek
President

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THE K-TRAN—made by Automatic Manufacturing Corporation—measures only ¾″ across. Yet it is available in RF and IF transformers covering frequency ranges from 20 KC to 30 MC and higher! For its size, it covers the widest range of uses in the IF field—and with unsurpassed stabilities. . . . As indicated, the makers credit K-TRAN's success, in large measure, to the controlled uniformity of G A & F Carbonyl Iron Powders.

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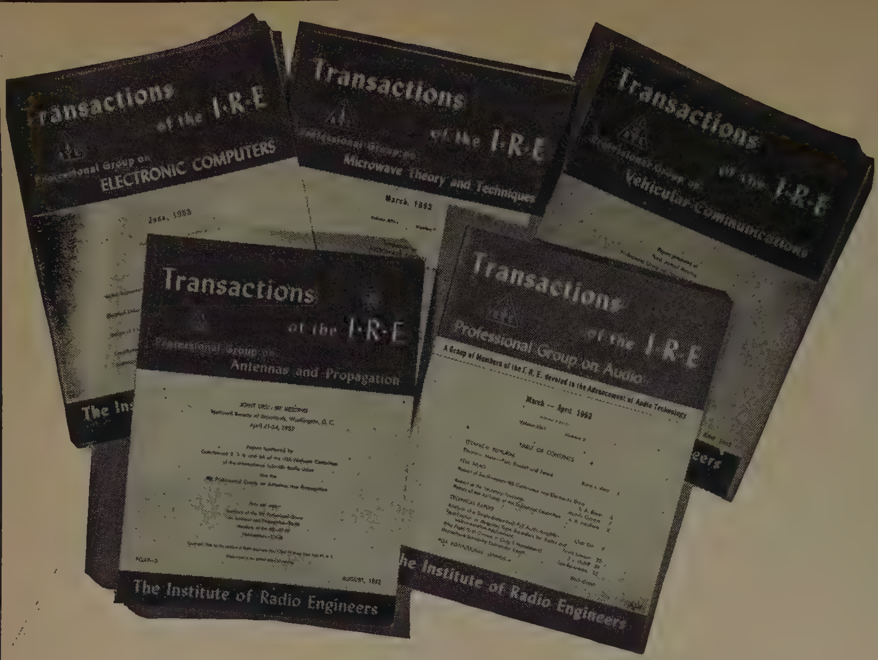
Transactions of the I·R·E
Professional Group on ELECTRONIC COMPUTERS
Transactions of the I·R·E
Microwave Theory and Techniques
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One of the most fundamental subjects in the radio engineering profession is circuit theory. It forms a basic part of every engineer's schooling. The methods of analyzing the characteristics of circuits and the techniques of designing circuits for optimum performance find wide application in practically every branch of communications and electronics. It is not surprising, therefore, that a substantial portion of the technical literature is devoted to this important subject.

While circuit theory is broad in its applications, the formulation of these methods and techniques is a sharply-defined and well-developed field in its own right, requiring specialized knowledge of a high degree. The IRE Professional Group on Circuit Theory is playing a key role in furthering the development of this field by making this specialized knowledge readily available to those who need it.

The Circuit Theory Group began operating in April, 1949 and since that time has become one of the largest and most active Groups in the IRE. Its most important activity has been the publication of its own technical periodical, called *Transactions*. Begun in 1952, the *Transactions* is issued quarterly to some 2500 members as a part of their two-dollar assessment. An interesting and valuable feature of *Transactions* is that many of the issues are built around a special topic. Among the topics thus treated or soon to be covered are Servo-mechanisms, Circuit Stability, Network Approximation, Nonlinear Filters, and Time-Variable Networks.

The Circuit Theory Group has been active also in sponsoring meetings and in organizing sessions at several national conferences. Numerous meetings are also held in several cities by local chapters of the Group. The net result has been to give engineers a golden opportunity to keep fully informed of the many significant developments in this important field.

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Chairman, Professional Groups Committee

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1947 First with "Penny Size" Servo Motors .938" diam. x 31/32" long.

1948 First with Servo Motors in Navy BuOrd Sizes 15 and 18 featuring straight-thru bores and integral ("potted")* stator-housing assemblies.

1949 First with Navy BuOrd Size 11 Synchros featuring "potted"* stator construction and straight-thru bores.

1950 First with Navy BuOrd Size 11 Servo Motors featuring "potted"* stator construction and straight-thru bores.

1951 First with Navy BuOrd Sizes 15 and 18 Servo Motor—Rate Generator combinations featuring "potted"* stator construction and straight-thru bores.

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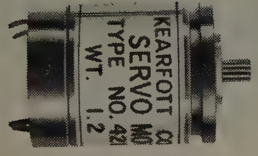
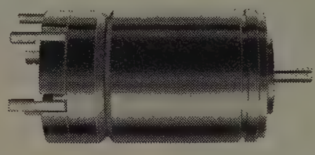
1953 First with Servo Motor-damping Generator combinations in Size 11 featuring "potted"* stator construction and straight-thru bores.

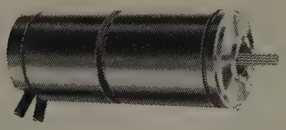
1954 First with 3/4" Diam. "Penny-Size" Servo Motors and Synchros. First with .05% linearity Tachometer Generators in Size 18. First with 115 volt Servo Motor Generator .980" in diameter. All Featuring "potted"* stator construction and straight-thru bores.



3/4" DIA. SERVO MOTOR—FULL SIZE



R500 SYNCHRO—3/4 SIZE



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Industrial Engineering Notes

(Continued from page 60A)

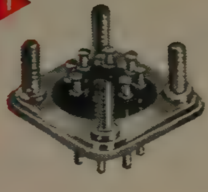
velopment of miniaturization techniques applicable to airborne military electronic equipment. The Publication, "Subminiaturization Techniques for Low-Frequency Receivers," is Circular 545, and can be obtained from the Government Printing Office, Washington 25, D. C., for 50 cents a copy. The work was carried out by the General Miniaturization Group, Engineering Electronics Section of the NBS, under support of the Bureau of Aeronautics, Navy Department. Earlier NBS work was reported in the Office of Technical Services publications, covering miniaturization as applied to high-gain fixed-tuned intermediate-frequency strips in the vhf band. This combines with the present developments to provide practical engineering treatment of a wide variety of electronic miniaturization problems over the radio frequency spectrum between 100 kc and 100 mc, the NBS announced. . . . **DYSEAC, a high-speed digital computer designed to serve as the experimental nucleus for a complex data-processing network, has been completed by the NBS.** The flexibility with which this machine controls and responds to a variety of external devices, which may include one or more full-scale computers of similar design, should enable scientists to explore diverse new areas of interest. Examples cited by the Bureau include the automatization of industrial and commercial operations, such as the "automatic factory" and the "automatic office," or any field where rapid information-processing and real-time control systems are necessary. Details were published in the September issue of *Tech. News Bull.*, NBS monthly publication. . . . **NBS also announced the development of a remote-control system which automatically measures radiation intensities and other variables in the vicinity of an atomic explosion and transmits the data by radio to a centrally located headquarters.** The system was designed at the request of the Atomic Energy Commission for use in nuclear tests. Though developed specifically for monitoring gamma radiation and weather conditions, it can be used with a wide variety of detectors to report many types of information. Except for the control station, the entire system is battery powered and will operate unattended for long periods. The radio link and the station interrogation system were designed and constructed by Motorola Inc., and are based largely on their $\frac{1}{2}$-w portable transmitter-receiver and other standard equipment. The radio link is a frequency-modulated system operating in the vhf band between 162 and 174 mc. Details on this new development were published in the August issue of the *Tech. News Bull.* . . . **The Atomic Energy Commission last week released 22 additional patents, including 11 in the electronics field. Nonexclusive, royalty-free licenses on the listed patents will be granted by the Commission as part of its program to make nonsecret technological information available**

(Continued on page 70A)

6

2

Get a head-start in mechanical assembly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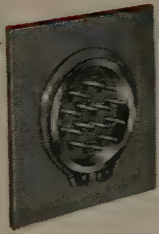
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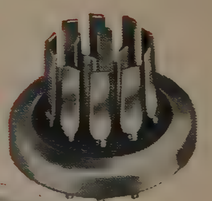
Let Hermetic's Vac-tite* Headers win part of the race for you with mechanical designs to solve your problems.

4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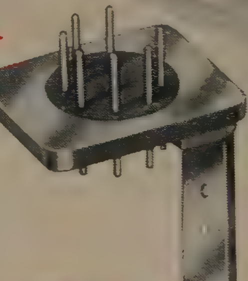
Here are just a few Hermetic Vac-Tite* seals that eliminate extra production operations and save you money!

1. **Unit Header with Studs Attached**—Saves space; shaped to fit enclosure or can; eliminates extra welding and soldering operations.
2. **Weld Seals**—Has the proper projections for leak-tight welds.
3. **Lock-Ring "Safety" Seal**—Simple, sure method for installing headers that is not dependent on solder alone for mechanical security; removable.
4. **Threaded Bushing Seal**—Firm mechanical connection has maximum shock and vibration resistance and adaptability for positioning and adjustment.
5. **Taper Tab Headers & Terminals**—Quick, solderless connections adaptable to many applications.
6. **Terminal Strip**—Pre-mounted terminals offer advantages of a conductive surface for heat dissipation, arc-resistance of glass, one piece assembly.
7. **Attached Bracket Seal**—Supports entire assembly on built-in structural member.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*VAC-TITE is HERMETIC's exclusive vacuum proof compression construction glass-to-metal seal.

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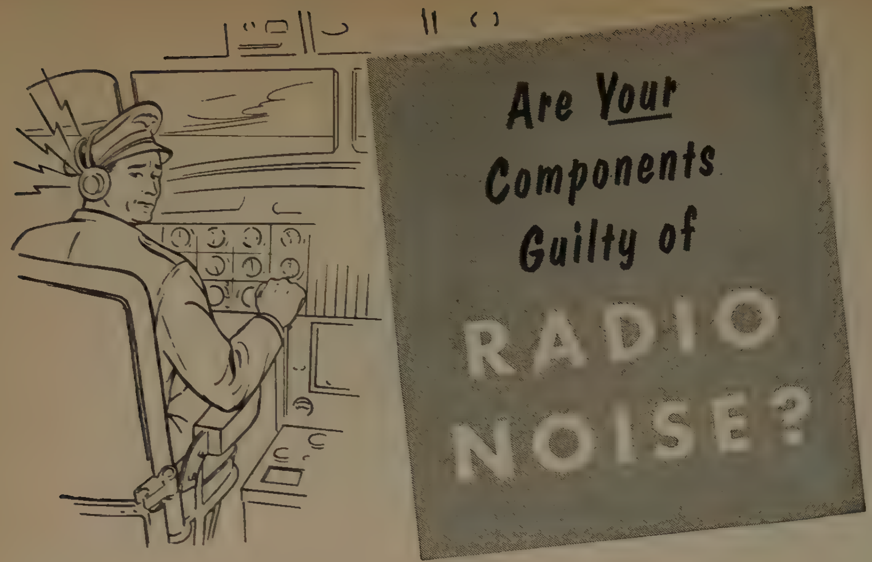


POTTER can tell you "why"

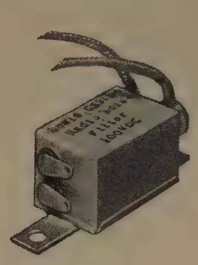
POTTER can tell you "how"

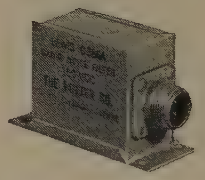
and **POTTER** can make
the FILTER that will
confirm that "how"

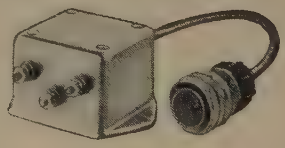
Once it's stated completely and correctly, a problem is half solved.

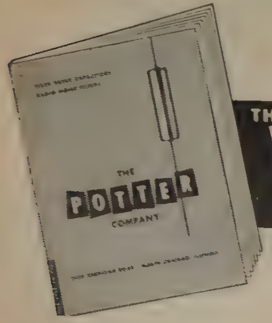
Potter can put the facts and figures of your problem on paper . . . can chart its limits in laboratory tests . . . can engineer the solution. And Potter can embody that solution in subsequent design and production.

Call Potter to engineer, design and produce the filter to solve your radio interference problem. Write for Bulletin 41E.









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Industrial Engineering Notes

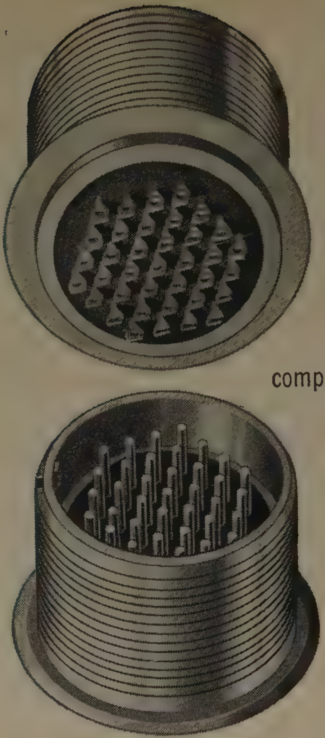
(Continued from page 68A)

for use by industry. Applicants for licenses should apply to the Chief, Patent Branch Office of the General Counsel, U. S. Atomic Energy Commission, Washington 25, D. C. identifying the subject matter by paten number and title. The following 11 patents of interest to the electronics industry were released: Phase Meter, 2,676,299; Electronic Scaling Circuits, 2,676,756; Ion Source, 2,677,061; Frequency Measuring Instrument, 2,677,104; Coincidence Circuit, 2,677,759; Ion Source, 2,677,771 Linear Cathode, 2,677,778; Head Seaming Device, 2,678,014; Ion Source, 2,679,597 Electrical Manipulator, 2,679,940, and Protective Circuit, 2,680,212. . . . **The Office of Technical Services, Commerce Department, has listed studies in the field of electronics in its June issue of the "Bibliography of Technical Reports." The following government-sponsored research reports can be purchased either from the Library of Congress or the Office of Technical Services, as indicated, and for the reported price:** "High Power Microwave Facilities of the RADC Research and Applied Techniques Laboratory" PB 113484, microfilm, $2.25; photostat, $4. Library of Congress, Photoduplications Section, Washington 25, D. C. "Investigation of Electronics Targets in the Prague Area" PB 113407, microfilm, $3; enlargement print, $9. Library of Congress. "Mechanized Production of Electronics" PB 111278, mimeo, $4. Office of Technical Services, Commerce Department, Washington 25, D. C. "On Adiabatic Amplification in the Microwave Region" PB 113571, microfilm, $2.25; photostat, $4. Library of Congress. "Beam-loading Effects in Small Reflex Klystrons" PB 113492, microfilm, $2.25; photostat, $4. Library of Congress. "Collinear Antenna Array: Theory and Measurements" PB 113574, microfilm, $4; photostat, $11.50. Library of Congress. "Electronic Admittance of the Reflex Klystron in the Presence of Space Charge" PB 113505, microfilm, $3.75; photostat, $10.25. Library of Congress. "Explicit Forms of the Propagation Function of a Traveling-Wave Tube" PB 113499, microfilm, $2.25; photostat, $4. Library of Congress. "Observations of Space Charge Phenomena in a Diode Magnetron" PB 113500, microfilm, $2.25; photostat, $4. Library of Congress. "Survey of Methods of Determining the Shunt Resistance for Klystron Resonant Cavities" PB 113497, microfilm, $3.25; photostat, $9. Library of Congress. "300–1000 Megacycle Traveling-Wave Tube Amplifier" PB 113503, microfilm, $3; photostat, $7.75. Library of Congress. "New Shop Techniques and Developments" PB 111338, mimeo, $1. Office of Technical Services. . . . **The OTS also has issued recently a new publication containing information on electronic capacitors.** The volume, compiled by the Wright Air Development Center especially for electronics design engineers, brings together a wealth of information on fixed capacitors from such widely scattered

(Continued on page 72A)

NOW 172 SERIES

a true HERMETIC SEAL connector



complete glass insert bead

individual glass contact bead

built-in "shock absorber!"

Thousands of tiny air bubbles act as an effective shock absorber in the improved glass seal of the new 172 series of Hermetic Seal Receptacles! Under a new manufacturing procedure which at last provides the electronics industry with a tough leak-proof hermetic connector, "hard" glass is heated to around 1800°F and cooled under compression. The glass assumes a cellular structure which has a leakage rate of zero and a strength which will withstand thousands of pounds of pressure per square inch!

corrosion-resistant surface!

A sealing treatment of the electro tin-plated shells of the 172 series receptacles gives them a surface which will resist salt-spray for a period of 100 hours! This sealing treatment also offers an excellent soldering surface. Connectors are available in individual glass contact bead and complete glass insert bead. They mate with standard AN plugs with female inserts.

For more information on the 172 series of Hermetic Seal Receptacles write and request Amform 2399

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RADIO INTERFERENCE
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measuring equipment

Stoddart NM-50A • 375mc to 1000mc

Commercial Equivalent of AN/URM-17

ULTRA-HIGH FREQUENCY OPERATION . . . Frequencies covered include UHF and color television assignments and Citizen's Band. Used by TV transmitter engineers for plotting antenna patterns, adjusting transmitters and measuring spurious radiation.

RECEIVING APPLICATIONS . . . Excellent for measuring local oscillator radiation, interference location, field intensity measurements for fringe reception conditions and antenna adjustment and design.

SLIDE-BACK CIRCUIT . . . This circuit enables the meter to measure the effect of the peak value of an interfering pulse, taking into account the shaping due to bandwidth.

QUASI-PEAK FUNCTION . . . An aid in measuring pulse-type interference, the Quasi-Peak function is just one of the many features of this specially designed, rugged unit, representing the ultimate in UHF radio interference-field intensity equipment.

ACCURATE CALIBRATION . . . Competent engineers "hand calibrate" each NM-50A unit. This data is presented in simplified chart form for easy reference.

SENSITIVITY . . . Published sensitivity figures are based on the use of the NM-50A with a simple dipole antenna or RF probe. However, the sensitivity of this fine instrument is limited only by the antenna used. The sensitivity of the NM-50A is better than ten microvolts across the 50 ohm input.

Stoddart RI-FI* Meters cover the frequency range 14kc to 1000mc

VLF
NM-10A, 14kc to 250kc
Commercial Equivalent of AN/URM-6B. Very low frequencies.

HF NM-20B, 150kc to 25mc
Commercial Equivalent of AN/PRM-1A. Self-contained batteries. A.C. supply optional. Includes standard broadcast band, radio range, WWV, and communications frequencies. Has BFO.

VHF
NM-30A, 20mc to 400mc
Commercial Equivalent of AN/URM-47. Frequency range includes FM and TV bands.

STODDART AIRCRAFT RADIO Co., Inc.
6644-C Santa Monica Blvd., Hollywood 38, California • Hollywood 4-9294

Industrial Engineering Notes

(Continued from page 70A)

sources as books, technical journals, and house organs. The publication, "Theory, Characteristics, and Applications of Fixed Electronic Capacitors," is available from the Office of Technical Services, Department of Commerce, Washington 25, D. C., for $1.25 per copy, and should be ordered by code number PB 111368. After introductory chapters on quality factors and general capacitor use, the volume takes up specific applications in such equipment as radio receivers, radio transmitters, and time-delay circuits. A valuable dictionary of capacitor applications and recommended types follows. Entries are arranged alphabetically by application—from audio amplifier through X-ray equipment. A miscellaneous section discusses such topics as capacitor inductance, three-terminal capacitors, buffer capacitors, dielectric absorption and impregnants. Sample problems relating to capacitor use and design are included, along with their solutions. A final section lists capacitor formulas and useful constants.

TELEVISION

The Federal Communications Commission on July 2, granted construction permits to two commercial television stations: Southwestern Publishing Co., Inc., Henderson, Nev., Channel 2 with 10.96 kw visual and 5.48 kw aural, and the Ohio Valley Broadcasting Co., Parkersburg, Clarksburg, W. Va., Channel 12 with 53 kw visual and 26.5 kw aural. Special temporary authorizations were granted to three television stations: WMSL-TV, Decatur, Ala., Channel 23; WLAC-TV, Old Hickory, Tenn., Channel 5, and KDRO-TV, Sedalia, Mo., Channel 6. One request to delete a previously authorized TV channel came from WHB-TV, Kansas City, Mo., Channel 9.

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are given in the Alphabetical Index to Radio-Electronic Manufacturers in the
New 1954
IRE DIRECTORY

custom transformers:

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the
swing
to

KEYSTONE?

KEYSTONE answers the needs of engineers and designers who have special transformer problems—gives them a recognized and established source for dependable quality.

KEYSTONE makes available special transformers of any type—400 cycle, plate filament or bias, saturable reactors (magnetic amplifiers), instrument, precision matched, and many others—with operating characteristics suited to any unusual or difficult specifications.

KEYSTONE transformers are tested under the most rigid conditions throughout production—can be qualified for approval under MIL-T-27 and other military and civilian specifications right in the plant, saving delay and costs.

When you have an unusual or difficult transformer application—when an ordinary transformer won't solve the problem—call on KEYSTONE for complete engineering and production to meet your *exact* requirements.

PERFORMANCE AND OPERATING CHARACTERISTICS OF THIS TYPICAL KEYSTONE TRANSFORMER:

driver transformer:
p-p plates to p-p grids
primary impedance:
15,000 ohms; Sec. imp. 940 ohms
freq. response:
± 3 db 20 cycles to 25 kc.
ratio: 4:1
balanced windings
pri. d.c.:
15 MA, D.C. Unbalance 1.5 MA
electrical center:
(pri & sec) ± .1%
harmonic distortion: less than 1%
mfg. in accordance with mil-t-27
size: 2 x 2 x 2 5/8

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It's easy to cut, shape or drill hard, brittle materials...
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The "Airbrasive" Unit can be operated manually or automatically. It's ideal for both laboratory and production line work.

INDUSTRIAL "AIRBRASIVE" PROCESS

With this revolutionary new cutting technique, a tiny stream of finely graded abrasive particles traveling at ultra high speeds does the work. There's no heat, no shock and no vibration. Consequently, the crystalline structure and other characteristics of the material remain unaffected. What's more, the process is fast and accurate and can be readily controlled.

Shown above is one of the many applications on which the "Airbrasive" process has been successfully used. In this case, the problem was to drill contact depressions .030″ in diameter and .015″ deep in a quartz disc. With the "Airbrasive" process, this was just another routine operation!

The "Airbrasive" process has solved many such "problem" jobs for electronics manufacturers—many of them considered impossible to do by conventional means. It has proved to be highly successful in cutting germanium and other hard, brittle materials—in "trimming" resistance elements on printed circuits—in removing deposited surface coatings—and in shaping fragile crystals used in neutron diffraction work.

Perhaps you have a similar problem. Why not arrange for a demonstration at our New York or California office. Or—if you prefer—we'll conduct tests on your samples and advise you as to the suitability of the process for your needs.

GET THE FACTS. BULLETIN 5307 has full information on where and how the "Airbrasive" process can be used. Send for your copy today.

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News—New Products

(Continued from page 33A)

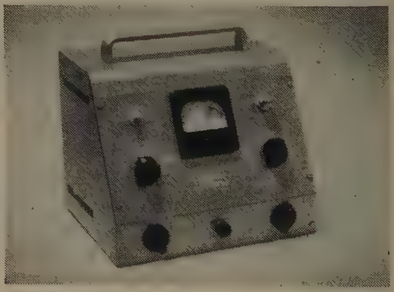
VTVM

The Model 1060 High Impedance Vacuum Tube Voltmeter developed by **Freed Transformer Co., Inc.,** 1715 Weirfield St., Brooklyn (Ridgewood), 27, N. Y., features: Input impedance of 50 megohms in parallel with 25 $\mu\mu f$ capacitor; Accuracy of 2 per cent on all ranges with full wave average reading meter calibrated in rms; and Frequency range 10 cps to 30 kc.

Specifications include voltage ranges of 0.001 volt to 100 volts in five ranges. Stability: Effect of variation in line voltage from 100 to 125 volts is less than 2 per cent, while effect in changes of tubes is less than ½ per cent. Scale: Logarithmic voltage scale calibrated from 1 to 10 plus a linear decibel scale calibrated from 0 to 20 db. Meter: 4 inch suppressed zero protected against overload.

Double-Purpose Power Supply

Suitable for a wide variety of laboratory power-supply applications, the new Spinco Duostat manufactured by **Specialized Instruments Corp.,** 589 O'Neill Ave., Belmont, Calif., includes, in a single unit, facilities for both constant-current operation and constant-voltage operation.



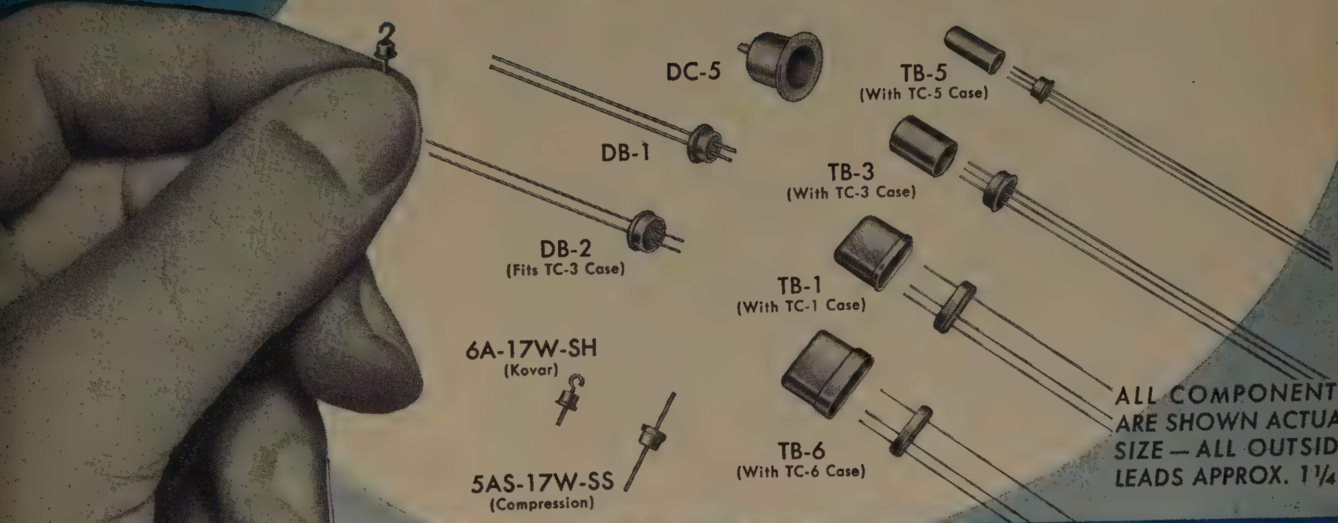
Energized from a standard 115-v 50/60 cps power outlet, the Duostat is capable of providing currents ranging from 5 to 50 milliamperes, automatically regulated to ±2 per cent at voltages ranging from 100 to 500 volts. In its constant-voltage mode of operation, the unit is self-regulated to ±1 per cent from 160 to 500 volts, and over a load-current range of 0 to 50 ma. Output is relatively unaffected by line-voltage changes between 105 and 125 volts.

Housed in an easily portable metal cabinet 9½ by 10¼ by 9½ inches, and having a top-mounted carrying handle, the Duostat weighs 11½ pounds.

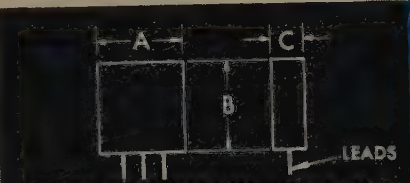
(Continued on page 76A)

For Hermetic Sealing

E-I MINIATURE TRANSISTOR COMPONENTS*


DC-5
TB-5
(With TC-5 Case)
DB-1
TB-3
(With TC-3 Case)
DB-2
(Fits TC-3 Case)
TB-1
(With TC-1 Case)
6A-17W-SH
(Kovar)
5AS-17W-SS
(Compression)
TB-6
(With TC-6 Case)

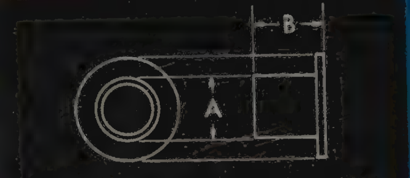
ALL COMPONENTS ARE SHOWN ACTUAL SIZE — ALL OUTSIDE LEADS APPROX. 1¼"



THREE ELECTRODE TYPE — Supplied complete with case. Suitable for transistor and other assemblies.

TYPE	A	B	C
TB-1	.335	.300	.183
TB-3	.217*	.300	—
TB-5	.132*	.325	—
TB-6	.398	.300	.191

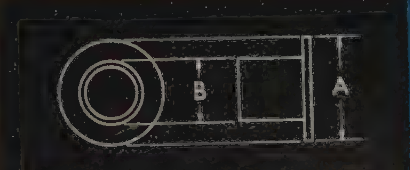
*Diameter



TWO ELECTRODE TYPE—Miniature headers for diode assemblies and similar applications.

TYPE	A	B
DB-1	.215	.170
DB-2	.217	.197
DC-5	.375	.275*

*With .050" O.D. and Projection for Resistance Welding.



SINGLE WIRE TYPE — Standard subminiature components available in compression and Kovar types.

TYPE	A	B
5AS	.125	.104
6A	.110	.088

These precision made components for transistors and other assemblies requiring hermetic sealing are available in three electrode, two electrode and single wire types. Where other than the standard types illustrated are required, E-I can custom design, and economically produce, special types in square, round or rectangular shapes. For complete information, call or write E-I — there is no obligation.

*PATENT PENDING
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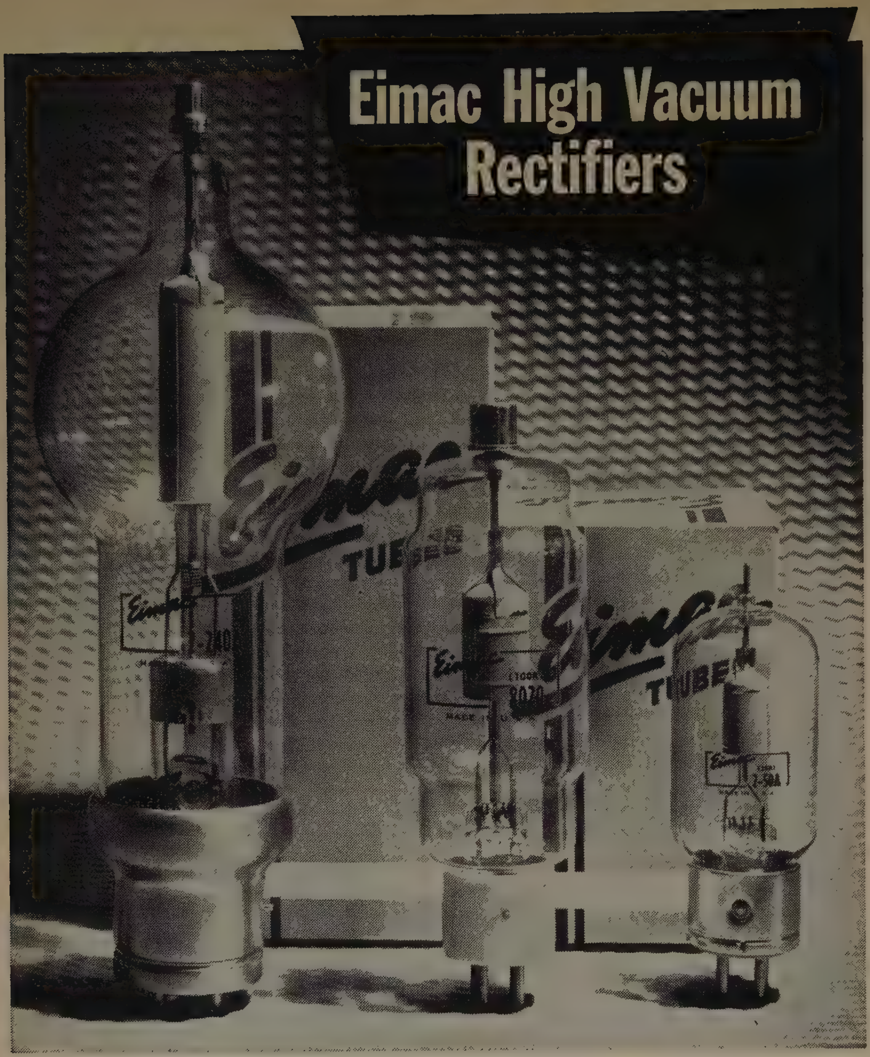


HIGH CURRENT, HIGH VOLTAGE OPERATION

Eimac's complete line of eight high vacuum rectifiers cover a wide range of average current, 50ma to 750ma and peak inverse voltages from 25,000v to 75,000v. In power supply units, voltage multipliers, pulse service or special applications at high frequencies, extreme ambient temperatures and high inverse voltages, Eimac high vacuum rectifiers are ideal. They give reliable performance at high frequencies and high voltages without generating radio frequency transients and have no lower limit to ambient operating temperature. Ruggedly constructed, Eimac high vacuum rectifiers contain many of the famous Eimac transmitting tube features such as an instant heating thoriated tungsten filament, that allows application of filament, plate voltages simultaneously; an exclusive radiation cooled pyrovac* plate; and elimination of internal insulators.

- *For additional information about Eimac high quality, high vacuum rectifiers, contact our Technical Services department.*

* An Eimac trade name.

EIMAC HIGH VACUUM RECTIFIERS

	PLATE			FILAMENT	
TYPE	Average Current MA	Dissipation Watts	Peak Inverse Voltage	Volts	Amps
2-25A	50	15	25,000	6.3	3.0
2-50A	75	30	30,000	5.0	4.0
8020	100	60	40,000	5.0	6.5
2-150D	250	90	30,000	5.0	13.0
250R	250	150	60,000	5.0	10.5
253	350	100	15,000	5.0	10.0
2-240A	500	150	40,000	7.5	12.0
2-2000A	750	1200	75,000	10.0	25.0

EITEL-McCULLOUGH, INC. SAN BRUNO CALIFORNIA
The World's largest manufacturer of transmitting tubes.

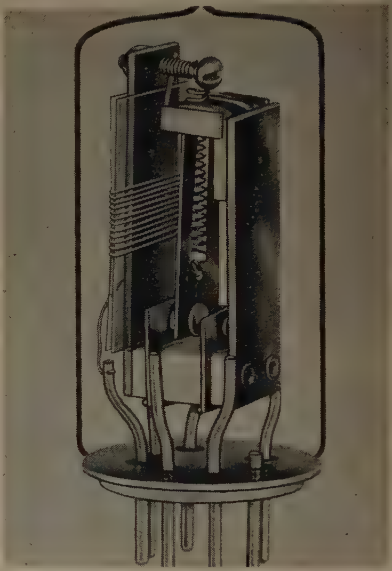
News—New Product

These manufacturers have invited PROCEEDING readers to write for literature and further technic information. Please mention your I.R.E. affiliatio

(Continued from page 74A)

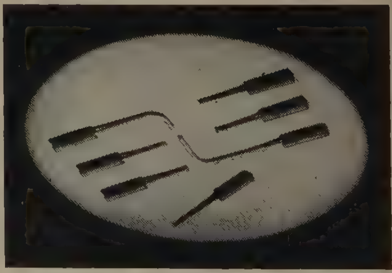
Thermal Relay

Elly Electronics Corp., P.O. Box 39 Fairlawn, N. J., has developed a therma time delay relay.



These Snapper relays have snap-action contacts in an inert gas-filled atmosphere. Time delay is from 2 seconds and up, the units are single-pole double throw, and are available in miniature or octal metal envelopes. They are designed for ambient temperature operation from −60 to +80°C. They operate on 6.3, 26.5 and 115 volts ac, or are built to requirements. The unit is capable of withstanding 30 G vibration at frequencies of from 5 to 55 cps.

New Ferrite Width and Linearity Cores



A new series of ferrite width and linearity cores for television receivers and industrial cathode-ray apparatus is now available from **Ferroxcube Corp. of America**, 235 East Bridge St., Saugerties, N. Y. Technical information is available upon letterhead request.

(Continued on page 107A)

1955
Radio Engineering Show
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General Ceramics

offering 3 complete services...

ENGINEERING-DESIGNING-PRODUCTION

on these 3 basic components...

ELECTRICAL INSULATORS
FERRAMIC® CORES
SOLDERSEAL TERMINALS

GENERAL
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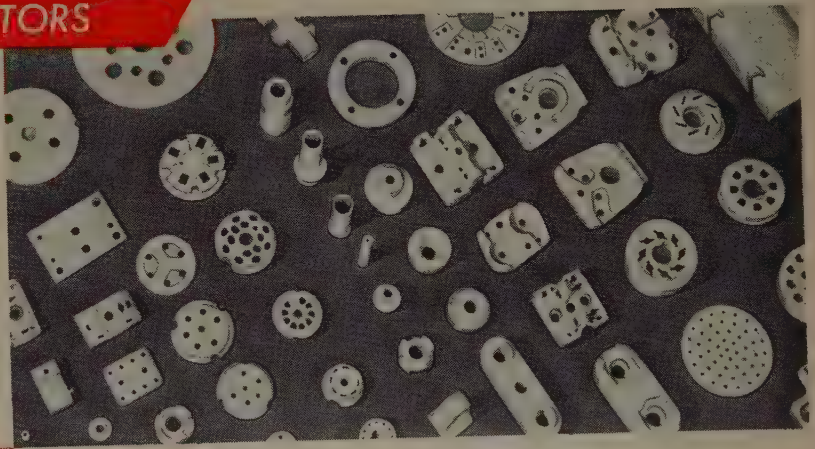
STEATITE INSULATORS

1. ELECTRICAL INSULATORS



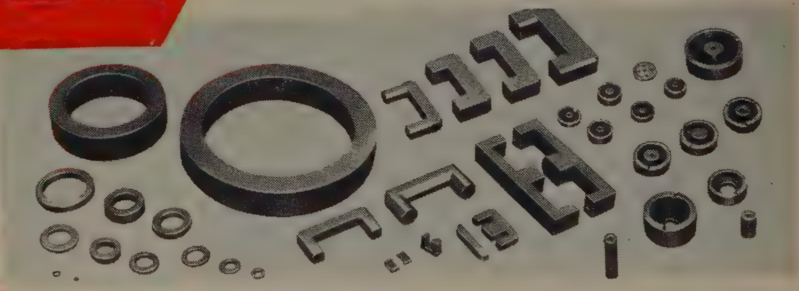
STEATITE—For low power loss at high frequency. High dielectric strength through wide temperature range. Low thermal expansion.

PORCELAINS—An economical high voltage material of great hardness. Low thermal expansion. Wet or dry process.

ALUMINA—Characterized by great hardness and chip resistance. Will withstand very high temperatures.

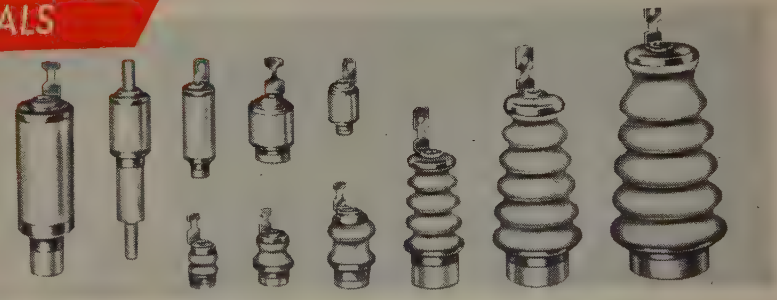
ZIRCON—Has low loss properties that vary inversely with frequency. An excellent high frequency material having good thermal shock resistance.

2. FERRAMIC CORES



General Ceramics Ferramic Cores are available in standard toroid, cup core and TV components. Standardization simplifies design problems, speeds delivery and lowers costs. The types illustrated are supplied in many grades of Ferramics for specific applications.

3. SOLDERSEAL TERMINALS



Featuring high mechanical strength, resistance to thermal shock and permanent hermetic sealing. Installation is easy and fast. Terminals are made of glazed Alumina Ceramic with lugs and eyelets hot tinned brass. Metallized areas are silver fired on ceramic, copper electroplated and tin fused for soft soldering.

For complete information on standard components, and recommendations on specific applications, call or write today; there is no obligation.

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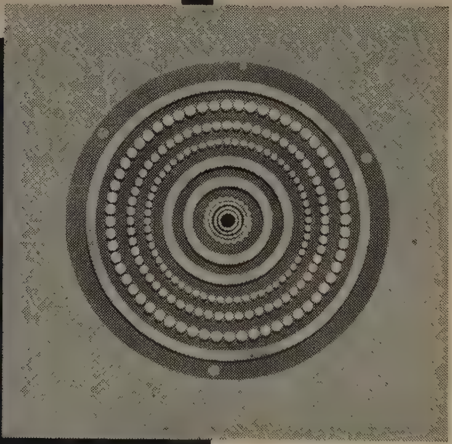

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Increased production of Mycalex 410 Telemetering Commutators now brings new savings to the user. Universally accepted as the finest commutator plate in the field, this precision-built unit injection-molded of Mycalex 410 glass-bonded mica insulation assures permanent dimensional stability – provides a tenacious bond to the metal inserts.

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- no dimensional change with age, humidity or temperature
- contact flatness within .0002"
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SINCE 1919 MYCALEX® THE INSULATOR

IRE People

Clifton H. Davis (M'53) has been recently appointed Assistant Chief Engineer of Neomatic, Inc., in Los Angeles, California. He will serve directly under T. Ross Welch, President and Chief Engineer, and will be concerned with current design engineering problems, new product research, and development work. Neomatic, Inc. designs, engineers and produces high precision sub-miniature electronic relays for application in advanced electronic equipment, computers, aircraft and guided missiles.

Clifton H. Davis

Mr. Davis has accumulated over 23 years of engineering experience in the radio and electronics field. Ten years of association with the U. S. Navy culminated in graduation from the Navy Radio Matériel School. He spent three years as Electronic Project Officer on such missiles as the Glomb, Azon, Rason, Felix, Gorgon, Bat, Loon and Lark and two years as civilian head of the Guidance Laboratory, U. S. Naval Missile Test Center, Point Mugu, California. For five years he was Chief Electronics Engineer, Globe Corporation, Aircraft Division, Joliet, Illinois. While with the Globe Corporation, Mr. Davis was responsible for the guidance and electronic components, instrumentation and flight testing of pilotless aircraft which included the KDG-2, KD2G-1, KD2G-2, KD4G-1, XKD5G-1, XKD6G-1 and KD6G-2 Target Aircraft. He was also charged with maintaining the Electronics and Instrumentation Laboratories.

⁂

Thomas J. Killian (A'47–SM'47–F'51), Chief Scientist of the Office of Ordnance Research, U. S. Army, has been appointed Dean of the School of Engineering and Architecture at the Catholic University of America, Washington, D. C. Assuming his new duties in September, he succeeds the Rev. Dr. Francis E. Fox, who died earlier this year.

Dr. Killian was educated at the Massachusetts Institute of Technology and at Princeton University. He taught at M.I.T. and later was dean of mathematics at Seattle University. Since 1941 he has served as engineer and scientist to both the Navy and the Army, and now holds the rank of Captain in the Naval Reserve.

Holding two American, a British, and a French patent, Dr. Killian is also contributor to many scientific journals. He is a member of Tau Beta Pi, Sigma Xi, and AIEE. He is a fellow of the American Physical Society and the American Association for the Advancement of Science, and a charter member of the IRE Committee on Nuclear Studies.

(Continued on page 82A)

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PROCEEDINGS OF THE I.R.E.

Stuart L. Bailey, Director, 1954–55 1476
The International Radio Consultative Committee .... H. Faulkner 1477
5066. Beam Deflection Color Television Picture Tubes .... James M. Lafferty 1478
5067. Correction to "Measurement of Resonant-Cavity Characteristics" .... G. L. Hall and P. Parzen 1495
5068. Characteristics of a Transmission Control Viewing Storage Tube with Halftone Display .... M. Knoll, H. O. Hook and R. P. Stone 1496
5069. IRE Standards on Electron Devices: Definitions of Semiconductor Terms, 1954 1505
5070. Measurement of Circuit Impedance of Periodically Loaded Structures by Frequency Perturbation .... E. J. Nalos 1508
5071. A Simple Microwave Correlator .... R. H. Wilcox 1512
5072. The Calibration of Amplitude Modulation Meters with a Heterodyne Signal .... J. Ross MacDonald 1515
5073. A New Transducer Diagram .... R. N. Bracewell 1519
5074. A Method of Making a Radar Self-Calibrating ...... E. A. Wolff 1521
5075. A Series Laminated Conductor for High Frequencies .... Hikosaburo Ataka 1527
5076. Analysis of Transmission Line Directional Couplers .... W. L. Firestone 1529
5077. The Power Spectrum of a Carrier Frequency Modulated by Gaussian Noise .... John L. Stewart 1539
5078. Step Discontinuities in Disk Transmission Lines .... R. N. Bracewell 1543
5079. Correction to "The Parallel-T Resistance-Capacitance Network" .... Laurence G. Cowles 1547
5080. Positive-Ion Trapping in Electron Beams .... E. L. Ginzton and B. H. Wadia 1548
5081. Impedance Measurement by Means of a Broadband Circular-Polarization Coupler .... Seymour B. Cohn 1554
5082. Weighting Functions for Time-Varying Feedback Systems .... J. A. Aseltine and R. R. Favreau 1559
5083. Analysis of Propagating Modes in Dielectric Sheets .... Leonard Hatkin 1565
5084. Correction to "Generalized Equations for RC Phase-Shift Oscillators" .... Sol Sherr 1568

Correspondence:

5085. "A Note on the Analysis of Vacuum Tube and Transistor Circuits" .... J. Shekel 1569
5086. "Rebuttal" .... L. A. Zadeh 1569
5087. "Finite Groups in Shannon Coding" .... D. A. Bell and T. C. Duggan 1569
Contributors to the PROCEEDINGS OF THE I.R.E. 1570

IRE NEWS AND RADIO NOTES SECTION

IRE News 1573
Seventh Annual Conference on Electrical Techniques in Medicine and Biology 1575
Radio Fall Meeting Program 1576
National Symposium of Quality Control and Reliability in Electronics 1577
Symposium on Fluctuation Phenomena in Microwave Sources 1577
Institute Committees 1580
Technical Committees 1581
5088. Abstracts of Transactions 1587
5089. Abstracts and References 1589
Meetings with Exhibits 2A
News and New Products 20A
Industrial Engineering Notes 36A
I.R.E. People 78A
Professional Group Meetings 86A
Section Meetings 90A
Student Branch Meetings 90A
Membership 93A
Positions Open 126A
Positions Wanted by Armed Forces Veterans 130A

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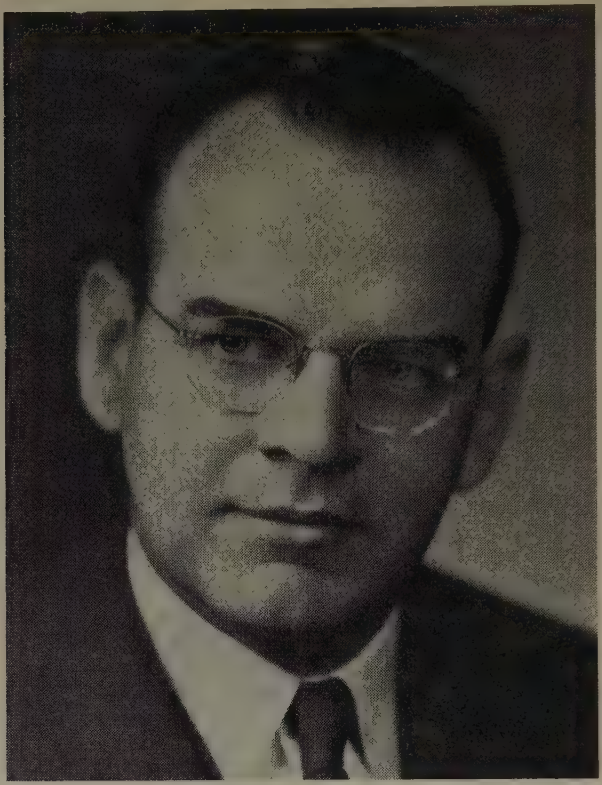

Stuart L. Bailey

DIRECTOR, 1954–1955

Stuart L. Bailey was born in Minneapolis, Minnesota, on October 7, 1905. He received the degree of Bachelor of Science in Electrical Engineering from the University of Minnesota in 1927 and the Master of Science degree from the same institution in 1928. During his undergraduate years he was active on the staff of W9X1, an experimental station run by the University, and while taking his graduate work he was chief engineer of WLB, owned and operated by the University of Minnesota. His master's thesis was on the subject of radio field-intensity measurement.

In the summer of 1928, Mr. Bailey became assistant radio engineer with the airways division of the United States Department of Commerce. His experience there included work on radio aids to marine and air navigation. He initiated and supervised the early work of the Lighthouse Service on radio-controlled fog signals and he also participated in the development of the visual radio range for use on the airways of the United States.

In the summer of 1929, Mr. Bailey went to Panama, where he installed two automatic marine radio beacons, one at the entrance to Cristobal Harbor and the other at Cape Mala, 120 miles south of Balboa.

In September 1930 he joined with C. M. Jansky, Jr., to form the consulting engineering firm of Jansky and Bailey. Mr. Bailey's activities in the consulting field have been on both general allocation problems and specific engineering guidance for broadcast stations and commercial operating companies. He has had charge of all of the laboratory activities of the firm.

During World War II, Mr. Bailey was in charge of all government-contract work done by the firm of Jansky and Bailey, a great deal of which was done under the Office of Scientific Research and Development. This work involved a detailed study of all factors affecting mobile short-range radio communication, a study of the effect of hills and trees as obstructions to radio transmission from 4 to 116 megacycles, and detailed analysis of electronic equipment to determine certain characteristics important to its operation by the armed services. In addition, under a Signal Corps contract, he supervised the work on measurement of many existing and proposed radio antennas for use by the ground forces.

Since June 1953, Mr. Bailey has been President of Jansky & Bailey, Inc., successors to the firm of Jansky & Bailey.

He became an Associate Member of The Institute of Radio Engineers in 1928, a Member in 1936, Senior Member in 1943, and was advanced to the grade of Fellow in 1943. He has been active on several IRE committees, and has served on the Board of Directors since 1943. He was a member of the Executive committee from 1945–1951, and has been a member now since 1953. He was appointed Treasurer of the IRE in 1948, and was elected President for the term 1949.

He is a member of Sigma Xi, Tau Beta Pi, and Eta Kappa Nu.

The International Radio Consultative Committee

H. Faulkner
Deputy Engineer-in-Chief
Post Office Engineering Department, London, England

There may be many members of the Institute of Radio Engineers and readers of its Proceedings who are not fully acquainted with the C.C.I.R. Since therefore the VIIth Plenary Meeting of this body was recently held in London, it may perhaps be an appropriate subject for discussion here.

The International Radio Consultative Committee (C.C.I.R.) is an organ of the International Telecommunication Union (I.T.U.), one of the Specialized Agencies of the United Nations. The I.T.U., whose job it is to regularize and facilitate international communications of all kinds, including radio, decided at its meeting in Washington in 1927 that in order to carry out its work efficiently it needed authoritative technical advice on radio matters and therefore set up the C.C.I.R., the first meeting of which was held at The Hague in 1929.

In addition to the ordinary Radio Administrative Conferences at which the various radio regulations, such as the allocation of frequency bands to the various services, the allocations of call signs, etc., are made, other Administrative Conferences are held from time to time at which plans are made assigning frequencies to individual stations for particular services such as broadcasting, marine, aeronautical, etc. The function of the C.C.I.R. is to formulate the technical basis on which these regulations and plans are built up. Thus the principal object is to study how the frequency spectrum can be used to the greatest advantage of all concerned, by reducing parasitic emissions from, and the bandwidths taken up by, radio transmitting stations, and by improving their frequency stability; by improving the frequency selectivity, sensitivity and stability of radio receivers, and by the use of directional aerials, etc.

Close study is also made of the propagation of radio waves of all frequencies and propagation curves are drawn from which service areas and interfering effects of transmitters can be calculated.

The work therefore can be viewed as of greatest importance to all radio communication users.

The C.C.I.R. meetings are attended by delegates and experts from countries all over the world, as well as by representatives of the major Radio Operating Agencies, whether Broadcasting, Radio, Telephone and Telegraph Fixed Services, Marine Services, or others, and thus become a clearing house for technical knowledge and experience from many diverse sources.

The work is divided into fourteen different Study Groups, the International Chairmen of which are responsible for gathering together all the information possible from their many National Correspondents between the meetings. When considered advisable the Chairmen call additional meetings of their Study Groups in between the Plenary Assemblies which are now due to meet every three years.

The Report of the Seventh Plenary Meeting which will shortly be published by the I.T.U. (Geneva, Switzerland) will contain all C.C.I.R. recommendations so far made, that are still valid, and will therefore be a valuable reference book for all radio engineers.

The principal value of the C.C.I.R. is that it brings to a common international point the latest information on good practice for radio services. If full use is made of this information and operators all over the world design and work their equipment accordingly, the maximum use will be made of that great asset of mankind, valuable beyond price, the radio frequency spectrum.

Beam Deflection Color Television Picture Tubes[*]

JAMES M. LAFFERTY[†], FELLOW, IRE

Summary—**The fundamental principles of color-switching at the screen by retarding or accelerating fields with a single electron beam are described. A mathematical analysis is made showing the effects and importance of various voltage and geometric parameters. The focusing effects of the apertures and the uniform fields are fully investigated. A new method of aperture-mask-voltage modulation is given for the accelerating-field case. Single gun tubes using these beam-deflection principles have the advantages of automatic color registry, simple monochrome presentation, short physical length, low color-switching electrode capacitance, and equal numbers of each of the three-color elements. Tubes using the retarding field are advantageous for large size screens, but ones using the accelerating field make much more efficient use of the electron beam. A method of computing aperture positions to give uniform variation in spacings is described. Some of the techniques described in the construction of 16- and 24-inch reflection-type tubes include prevention of microphonics in the screen; printing methods that make webbing lines invisible; glass spacer blocks that withstand high voltage gradients; a method of etching slits in the screen at an angle to reduce beam-current interception; screen alloys that withstand high-power dissipation; and a high-current electron gun.**

INTRODUCTION

THE METHOD of switching colors to be discussed here is one in which the color-control mechanism is located in the vicinity of the phosphor screen. In its simplest form, only one electron beam is required; however, multiple beams may be employed. The electron beam is deflected by means of an electric field. The incident beam is directed so that its initial velocity always has a component transverse to the electric field.

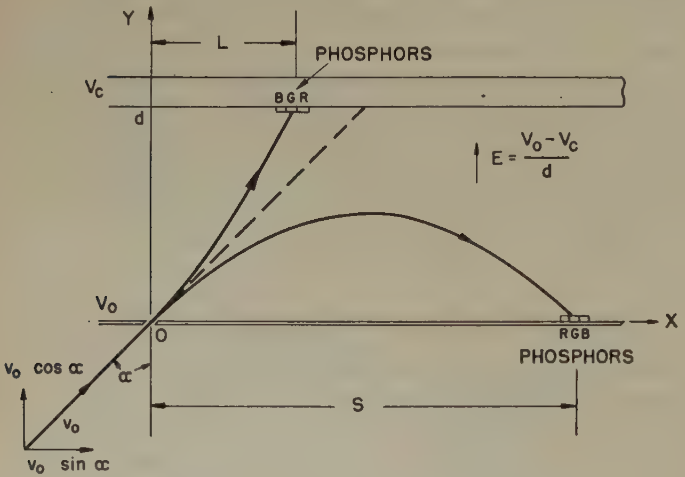

Fig. 1—Electron trajectories in a uniform field.

Such an arrangement is shown in Fig. 1. A uniform electric field E is produced between two parallel-plane electrodes at potentials V_0 and V_c and separated a distance d. An electron beam entering this field through an opening in one of the electrodes has a component of velocity, $v_0 \sin \alpha$, normal to the field. The disposition of the beam in this field will depend on the magnitude and direction of the field with respect to the velocity of the beam. If $V_0 = V_c$, the electric field is zero, and the beam will continue on in a straight line, striking the upper electrode. If $V_c > V_0$, the electric field is negative, and the beam will be accelerated towards the upper electrode, following a parabolic path that is concave upward. If $V_c < V_0$, the electric field is positive, and the beam will follow a parabolic path that is concave downward. It will be shown further that, if $V_c > V_0 \sin^2 \alpha$, the beam will still strike the upper electrode, and if $V_c < V_0 \sin^2 \alpha$, the beam will return to the lower electrode. There are, thus, two modes of operation for this device; one in which the beam strikes the upper electrode and the other in which it strikes the lower electrode.[1] The exact point at which the beam strikes either of the electrodes may be controlled by varying the velocity (in magnitude and/or direction) of the electron beam or the electric field, or a combination of both. This, then, offers the possibility of having the electron beam excite a series of color phosphors in a sequential manner by variation of the proper parameters.

ELECTRON BALLISTICS

In order to determine the voltage required for switching colors, the screen configuration, and the limits on various tolerances, it is necessary to determine various mathematical relations between the electron motion, the applied voltages, and the tube geometry.

The location of the Cartesian co-ordinates with respect to the electrodes is shown in Fig. 1. It will be assumed here that the space-charge effects are negligible. An electron beam from a gun with its cathode at potential V_k is directed through a flat aperture mask which is at a potential of V_0 volts, by means of a conventional scanning system. Parallel to the mask, and spaced a distance d from it, is the transparent conducting electrode at a potential V_c. These voltages produce a uniform field $E = (V_0 - V_c)/d$ in the direction of the y-axis between the electrodes. The electron beam enters the field, making an angle of incidence α with the screen.

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[*] Decimal classification: R583.6×R138.312. Original manuscript received by the IRE, May 24, 1954; revised manuscript received, August 6, 1954.

[†] General Electric Res. Lab., Schenectady, N. Y

[1] A tube in which the beam strikes the lower electrode has been previously described in detail. See P. K. Weimer and N. Rynn, "A 45-degree reflection-type color kinescope," PROC. I.R.E., vol. 39, pp. 1201–1211; October, 1951.

It can easily be shown[2] that the electron trajectory equation which applies between the electrodes is

$$y = x \cot \alpha - x^2(1 - V_c/V_0)/4(1 - V_k/V_0)d \sin^2 \alpha. \quad (1)$$

By letting $V_k = 0$, differentiating (1) with respect to x, and setting the results equal to zero, it is found that $x = (d \sin^2 \alpha)/(1 - V_c/V_0)$. Substituting this value back in (1) gives the ordinate of the vertex of the parabolic trajectory:

$$y_m = (d \cos^2 \alpha)/(1 - V_c/V_0). \quad (2)$$

If $y_m = d$, then, on solving (2), $V_c/V_k = \sin^2 \alpha$ and the beam is tangent to the upper electrode. If $V_c/V_0 < \sin^2 \alpha$, the beam will be reflected and return to the lower electrode, and if $V_c/V_0 > \sin^2 \alpha$, the beam will strike the upper electrode.

The displacement of the beam, i.e., the horizontal distance S or L (Fig. 1) between the point where the beam enters the electric field and the point where it strikes one of the electrodes, is found by setting y in (1) equal to zero or d, respectively, and solving for x. These two cases will be considered separately.

Reflection Case, $V_c/V_0 < \sin^2 \alpha$

Setting (1) equal to zero and solving for x gives:

$$S = 2d \frac{1 - V_k/V_0}{1 - V_c/V_0} \sin 2\alpha. \quad (3)$$

Let

$$d_0 = d/(1 - V_c/V_0), \quad (4)$$

and, if $V_k = 0$, then (3) becomes

$$S = 2d_0 \sin 2\alpha. \quad (5)$$

The physical significance of d_0 may be seen in Fig. 1. It is the distance, measured parallel to the y-axis, from the mask to the zero potential plane in the electric field. If V_c is negative, the zero potential plane lies between the two electrodes. If V_c is positive, the zero potential plane is a virtual plane located outside the electrodes at a distance that can be determined by assuming that the electric field extends beyond the electrodes with the same value it has between them.

From (5) it can be seen that the displacement of a beam that has an angle of incidence of 45 degrees is just twice the distance from the screen to the zero potential surface. This displacement will be called S_0. If the angle of incidence is made greater or less than 45 degrees, the displacement drops off symmetrically about the 45-degree value.

The displacement of the returning electron beam may be varied by changing the potential of the cathode, mask or the reflector (upper) electrode. The rate of change of displacement of returning beam with respect to cathode voltage is found by differentiation of (3):

$$\frac{dS}{dV_k} = -\frac{2d(1 - V_k/V_0) \sin 2\alpha}{V_0(1 - V_c/V_0)}. \quad (6)$$

If $V_k = 0$, this reduces to

$$\frac{dS}{dV_k} = -\frac{S}{V_0}. \quad (7)$$

The effect of changes in beam or mask voltage for $V_k = 0$, may be found by differentiation of (5) with respect to V_0:

$$\frac{dS}{dV_0} = -\frac{SV_c}{(1 - V_c/V_0)V_0^2}. \quad (8)$$

On modulating the reflector voltage, the rate of change of displacement of the beam is found by differentiation of (5):

$$\frac{dS}{dV_c} = \frac{S}{V_0(1 - V_c/V_0)}. \quad (9)$$

A comparison of (7), (8), and (9) shows that

$$\frac{dS}{dV_0} = -(V_c/V_0)\frac{dS}{dV_c} = \frac{V_c/V_0}{1 - V_c/V_0}\,\frac{dS}{dV_k}. \quad (10)$$

If the reflector voltage is zero, the displacement of the beam is independent of beam voltage. If the cathode voltage and the reflector voltage are both equal to zero, then (7) and (9) reduce to the same value except for sign. This offers the possibility in certain applications of switching color by modulating the cathode voltage.

Transmission Case, $V_c/V_0 > \sin^2 \alpha$

The distance L which the beam is displaced along the screen in the x direction is determined by setting $y = d$ in (1) and solving for x:

$$L = \frac{(1 - V_k/V_0)d \sin 2\alpha}{1 - V_c/V_0} \cdot \left[1 - \left(\frac{V_c/V_0 - \sin^2 \alpha - (V_k/V_0)\cos^2 \alpha}{(1 - V_k/V_0)\cos^2 \alpha}\right)^{1/2}\right]. \quad (11)$$

If the cathode is at zero potential, this expression reduces to

$$L = \frac{d \sin 2\alpha}{1 - V_c/V_0}\left[1 - \frac{(V_c/V_0 - \sin^2 \alpha)^{1/2}}{\cos \alpha}\right]. \quad (12)$$

Under these conditions if $V_c/V_0 < \sin^2 \alpha$, the electron beam will fail to reach the screen, and L becomes imaginary.

The beam displacement L may be varied by changing the cathode, mask, or screen voltages. The rate of change of beam displacement with respect to the cathode voltage is found by differentiation of (4):

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[2] J. R. Pierce, "Theory and Design of Electron Beams," D. Van Nostrand Company, Inc., New York, N. Y., 1st ed., p. 21, eq. (3.11); 1949.

$$\frac{dL}{dV_k} = -\frac{d \sin 2\alpha}{(1 - V_c/V_0)V_0}\left[1 - \frac{V_c/V_0 + \cos 2\alpha - 2(V_k/V_0)\cos^2\alpha}{2[V_c/V_0 - \sin^2\alpha - (V_k/V_0)(1 - V_k/V_0)\cos^2\alpha]^{1/2}\cos\alpha}\right]. \quad (13)$$

For the case of zero cathode potential, this reduces to

$$\frac{dL}{dV_k} = \frac{-d \sin 2\alpha}{(1 - V_c/V_0)V_0}\left[1 - \frac{V_c/V_0 + \cos 2\alpha}{2[V_c/V_0 - \sin^2\alpha]^{1/2}\cos\alpha}\right]. \quad (14)$$

The rate of change of beam displacement with respect to the screen voltage for $V_k = 0$ is found by differentiation of (5):

$$\frac{dL}{dV_c} = \frac{d \sin 2\alpha}{(1 - V_c/V_0)^2 V_0}\left[1 - \frac{V_c/V_0 + \cos 2\alpha}{2[V_c/V_0 - \sin^2\alpha]^{1/2}\cos\alpha}\right]. \quad (15)$$

Differentiation of (5) with respect to V_0 gives the rate of change of beam displacement with respect to the beam or mask voltage for $V_k = 0$.

$$\frac{dL}{dV_0} = \frac{-d(V_c/V_0)\sin 2\alpha}{(1 - V_c/V_0)^2 V_0} \cdot \left[1 - \frac{V_c/V_0 + \cos 2\alpha}{2[V_c/V_0 - \sin^2\alpha]^{1/2}\cos\alpha}\right]. \quad (16)$$

A comparison of (14), (15), and (16) shows that

$$\frac{dL}{dV_0} = -(V_c/V_0)\frac{dL}{dV_c} = \frac{V_c/V_0}{1 - V_c/V_0}\frac{dL}{dV_k}, \quad (17)$$

which is identical to (10) for the reflection case.

FOCUSING ACTION

The electric field has two focusing effects on the beam. The first occurs at the point where the beam enters the field through the aperture in the mask, and the second occurs throughout the trajectory as the beam traverses the field.

The first focusing effect is caused by apertures in the mask acting as lenses by virtue of the difference in electric fields on the two sides of the mask. If the openings are in the form of slits, the focal length[3] is given approximately by

$$f = -(2V_0 \cos^2\alpha)/E. \quad (18)$$

If the value of the electric field E is substituted in this expression, it reduces to

$$f = \frac{-2d \cos^2\alpha}{1 - V_c/V_0} = -2d_0 \cos^2\alpha. \quad (19)$$

If $V_c < V_0$, the electric field is in the direction of the y-axis. This retarding field causes the slit to act as a diverging lens with a negative focal length. If $V_c > V_0$, the accelerating field produces a converging lens action at the slit.

In the second focusing effect a retarding field between the mask and reflector plate produces a focusing action within the field in the plane of incidence of the beam. This may be seen by differentiation of (5):

$$dS = 2S_0 \cos 2\alpha d\alpha. \quad (20)$$

Electrons in a beam traveling with uniform velocity that are diverging from the axis of the beam with an angle $d\alpha$ will, after traveling a distance S_0, be displaced transversely a distance $dS = S_0 d\alpha$. If this criterion is applied to (20), it can be seen that, if 2 cos 2α is less than 1, a focusing action occurs, and if this quantity is greater than 1, a defocusing action occurs. Thus, greatest focusing action, due to the retarding field, occurs when the angle of incidence is 45 degrees, and defocusing occurs when it is less than 30 degrees or greater than 60 degrees.

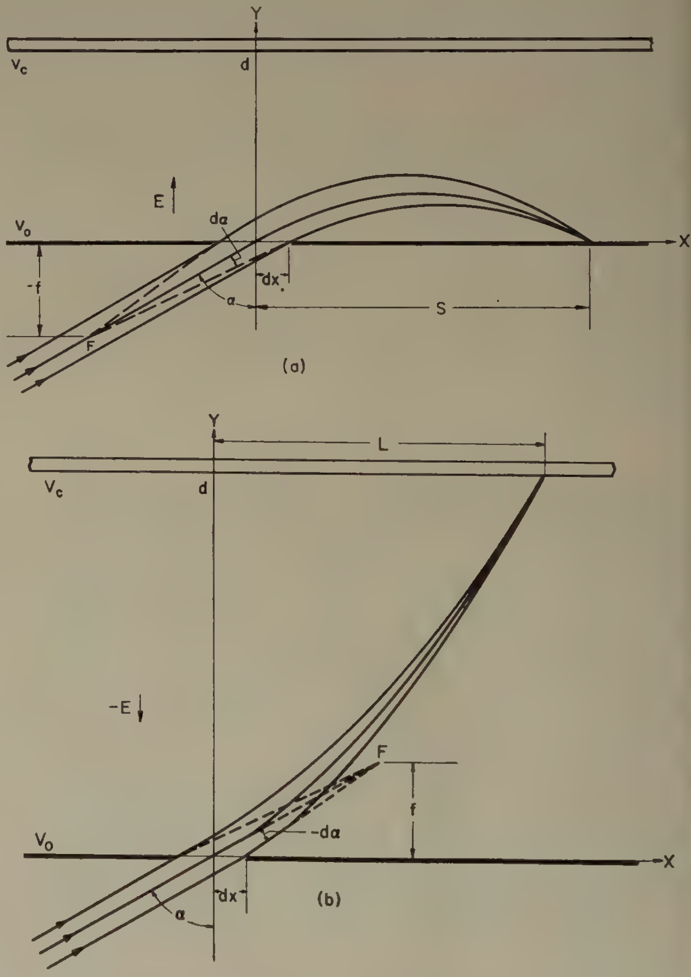

Fig. 2—Focusing action in (a) a retarding field, and (b) an accelerating field, taking into account the lens field at the slit.

Considering again the aperture lens effect, Fig. 2(a) shows an initially parallel beam passing through a slit of width 2 dx at an angle of incidence α, into a retarding field E. Electrons that pass through the center of the slit have their direction unaltered, while those that pass near the edge of the slit are given an angular divergence $d\alpha$, so that the beam appears to come from the focal point F. From the geometry of Fig. 2(a), it is seen that $d\alpha = dx \cos \alpha/(-f/\cos a) = -(dx \cos^2 \alpha)/f$. Substituting the value of the focal length from (19) in this expression gives $d\alpha = dx/S_0$. On combining this with the field focusing effect by substitution in (20), it is found that

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[3] This is a more general form of the well-known Davisson and Calbrick formula. It is derived in the Appendix.

$$dS = 2 \cos 2\alpha dx. \tag{21}$$

If the electrons which pass through the lens near the edge of the slit at $x = dx$ are to strike the screen at the same point as those which pass through the center of the lens, their displacement must be shorter than those which pass through the center by the amount dx. Thus, for perfect focusing dS must equal $-dx$. When this substitution is made in (21), it is found that $\alpha = 60$ degrees. Thus, optimum line-focusing occurs when the beam passes through the aperture mask at a 60-degree angle of incidence, as shown in Fig. 2(a). For angles greater than 60 degrees, over-convergence occurs, and for angles less than 60 degrees the beam fails to completely converge. However, for angles greater than 45 degrees and less than 90 degrees, the width of the beam on returning to the mask will be less than the width of the slit. From the standpoint of focusing, the sweep angle should be centered about a 60-degree angle of incidence. Centering the sweep angle about a 45-degree angle of incidence, as suggested by Weimer and Rynn,[4] will give an image whose width will be less susceptible to variations in the angle of incidence, but will not give an image of minimum width.

If circular apertures are used in the mask in place of the slits, the focal length of the aperture is doubled and the beam fails to completely converge in the retarding field for all angles of incidence less than 90 degrees. However, for angles greater than 45 degrees, the returning beamwidth will be less than the diameter of the aperture in the plane of the beam.

Although it has no direct bearing on the problem at hand, it is of interest to examine the dimensions of a returning beam which enters parallel to the retarding field. From (21), if $\alpha = 0$, then $dS = 2dx$. Hence, the width of the beam will be three times the width of the slit. Similarly, for a circular aperture, the returning beam diameter is twice the diameter of the aperture.

For accelerating field case in which $V_c/V_0 > \sin^2 \alpha$, differentiation of (12) with respect to α gives

$$dL = 2d_0\{\cos \alpha[2 \cos \alpha - (V_c/V_0 - \sin^2 \alpha)^{1/2} + \sin^2 \alpha(V_c/V_0 - \sin^2 \alpha)^{-1/2}] - 1\} d\alpha. \tag{22}$$

Fig. 2(b) shows an initially parallel beam passing through a slit of width $2dx$ at an angle of incidence α into an accelerating field. On passing through the slit, the beam is focused toward the focal point F. From the geometry of Fig. 2(a), $-d\alpha = (dx \cos \alpha)/(f/\cos \alpha) = (dx \cos^2 \alpha)/f$. On substituting the value of the focal length from (19), this expression becomes $d\alpha = dx/2d_0$. Substituting this in (22) and letting $dL = -dx$ for a line focus, it is found on solving for V_c/V_0 that

$$V_c/V_0 = 2(1 + \cos \alpha). \tag{23}$$

Thus, for any particular angle of incidence α, a sharp line-focus may be obtained on the screen by adjusting the voltage ratio V_c/V_0 to satisfy (23). When the incident beam is parallel to the field, the screen voltage must be four times the beam voltage to obtain line-focus.

If the focus condition (23) is substituted in (22), it is found that

$$dL = d\left(\frac{2}{1 + 2 \cos \alpha}\right) d\alpha. \tag{24}$$

This equation shows that if V_c/V_0 is always adjusted to give optimum focus, $dL/d\alpha$ increases slowly with the angle of incidence from $(2/3)d$ at $\alpha = 0$ degrees to $2d$ at 90 degrees. If the focus criterion that was applied to (20) is used here, and if d plays the same role as S_0 in (20), it will be seen that a focusing action occurs when α is less than 60 degrees and a defocusing action occurs when α is greater than 60 degrees.

It is of interest to note that when (23) is substituted in (12), (15), and (16), they simplify to

$$L = \frac{2d \sin \alpha}{1 + 2 \cos \alpha}, \tag{25}$$

$$\frac{dL}{dV_c} = -\frac{d \sin \alpha}{(1 + 2 \cos \alpha)^2(1 + \cos \alpha)}, \tag{26}$$

and

$$\frac{dL}{dV_0} = \frac{2d \sin \alpha}{(1 + 2 \cos \alpha)^2 V_0}. \tag{27}$$

MODULATION METHODS FOR THE TRANSMISSION TUBE CASE

From (23) it is found that for, $\alpha = 60$ degrees, optimum post-acceleration focusing occurs when the screen voltage is 3 times the beam voltage. If this substitution is made, (17) gives

$$\frac{dL}{dV_0} = -3\frac{dL}{dV_c} = -1.5\frac{dL}{dV_k}. \tag{28}$$

Thus it is seen that, of the three methods of modulation, the deflection sensitivity is least for the screen-voltage modulation and greatest for the aperture-mask-voltage modulation. The high sensitivity of this latter method results from the favorable combination of two effects. In Fig. 1, L may be increased by decreasing the electric field E or by increasing the initial velocity of the electron beam as it enters this field. Increasing the beam voltage does both of these things simultaneously, while changing the screen or cathode voltage does only one of these at a time.

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[4] Weimer and Rynn, loc. cit.

Holding the screen and anode (aquadag wall coating) voltages constant, and modulating the mask voltage, would also simplify the shielding problem for high color-switching rates. By keeping the wall coating at a constant potential, the deflection sensitivity (at the yoke) will not change on modulating the aperture mask. There will, however, be a weak lens field established between the wall coating and the mask. This will cause a slight displacement and change in the angle of incidence of the beam as it strikes the aperture mask. Calculations made by A. B. Brown, Jr., of this laboratory, show that for a practical case this displacement may be the width of several raster lines and that the change in angle of incidence may be as much as 2 degrees. By placing a grid which is at the aquadag-coating potential near the mask, the fringing field may be confined and the beam displacement at the mask made negligible, while the change in angle of incidence remains the same. Under these conditions, modulation of the mask voltage will change the displacement of the beam at the screen because of three things: (1) change in the angle of incidence α; (2) change in the initial velocity of the beam as it enters the accelerating field; (3) change in the accelerating field. This may be written

$$dL = \frac{\partial L}{\partial \alpha}\, d\alpha + \frac{\partial L}{\partial V_0}\, dV_0$$

or

$$\frac{dL}{dV_0} = \underset{(a)}{\frac{\partial L}{\partial \alpha}}\ \underset{(b)}{\frac{d\alpha}{dV_0}} + \underset{(c)}{\frac{\partial L}{\partial V_0}} \cdot \tag{29}$$

Terms (a) and (c) in this equation are given by (24) and (27), respectively. Term (b) may be found as follows. If the beam makes an angle of incidence α_0 with the grid which is at potential V_0', then by application of (1) it may be shown that the angle of incidence that the beam has on arrival at the mask is given by

$$\tan \alpha = (V_0/V_0' - \sin^2 \alpha_0)^{-1/2} \sin \alpha_0. \tag{30}$$

Solving this expression for V_0 and differentiating with respect to α gives $dV_0/d\alpha = +2V_0' \sin^2 \alpha_0 \cot \alpha \csc^2 \alpha$. Since $V_0' \approx V_0$ and $\alpha \approx \alpha_0$, this reduces to

$$\frac{d\alpha}{dV_0} = -\frac{\tan \alpha}{2V_0} \tag{31}$$

thus giving term (b) for (29). Making the substitutions in (29) gives

$$\frac{dL}{dV_0} = \frac{d}{V_0}\left[-\frac{\tan \alpha}{1 + 2\cos \alpha} + \frac{2 \sin \alpha}{(1 + 2\cos \alpha)^2}\right]. \tag{32}$$

The first term of this equation, which gives the change in displacement of the beam due to the change in the angle of incidence, is always larger in magnitude than the second term and causes the beam displacement to change in the direction opposite to that given by the second term. At $\alpha = 60$ degrees, the first term is just twice the second, so that the beam displacement is changed the same amount as given by (27), but in the opposite direction. Hence, for values of α less than 60 degrees, (32) gives changes in displacement that are less in magnitude than (27); and for α greater than 60 degrees, (32) gives values that are greater than (27). Eq. (32) is plotted in Fig. 3. Even at the expense of adding an extra grid, there would appear to be considerable advantage in switching colors by modulating the aperture-mask voltage.

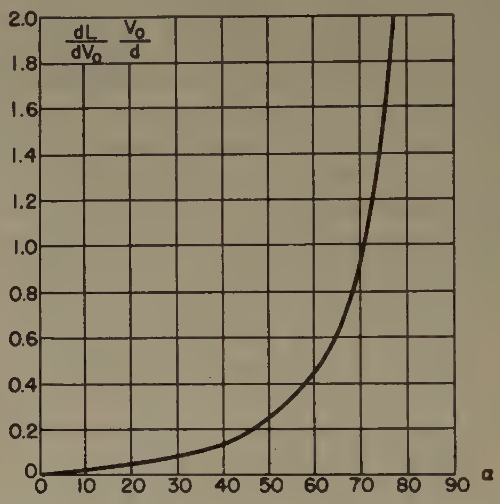

Fig. 3—Beam-deflection sensitivity as a function of the angle of incidence for aperture plate modulation.

An outline drawing of the proposed tube is shown in Fig. 4. It should be pointed out that an angle of incidence of 45 degrees has no significance for this tube, and the greater the angle of incidence of the electron beam with the mask, the greater will be the deflection sensitivity. How far one can go in this direction depends on a compromise of adequate focusing, sweep and keystone correction. The color-switching voltage for the case $\alpha = 60$ degrees and $V_c/V_0 = 3$ is, according to Fig. 3:

$$\Delta V_0 = \frac{\Delta L V_0}{0.433 d} \cdot \tag{33}$$

As an example, consider the case in which $V_c = 20$ kv, $V_0 = 6.67$ kv, $d = 1/2$ inches and $\Delta L = 0.015$ inches (phosphor-stripe spacing). Substituting these values in (33) gives 462-volts peak to switch from one color stripe to the next. Complete color-switching then requires a sine wave voltage of 327-volts rms on aperture mask.

The capacitance of a 20-inch rectangular aperture mask for this case would be approximately 175 $\mu\mu$fd.

Thus, the power required to modulate the mask would not be excessive.

It will be noted in Fig. 3 that the beam-deflection sensitivity changes rapidly with angle of incidence. This is undesirable from the standpoint of obtaining a screen with uniform phosphor-stripe spacing. This difficulty may be corrected to a large extent by tilting the screen with respect to the aperture mask, so that the spacing is increased for small angles of incidence and decreased for large angles of incidence. This difficulty may be even further improved by curving the aperture mask and screen to give a nearly constant angle of incidence for all points on the mask.

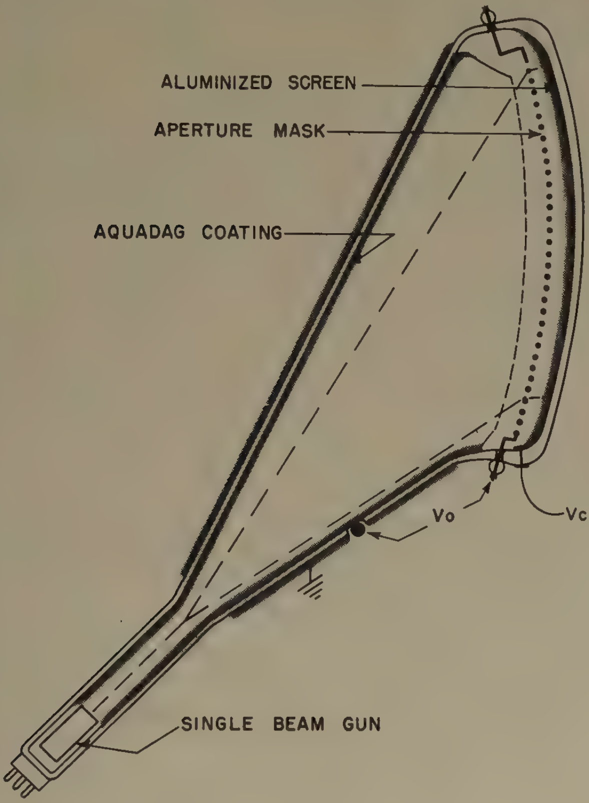

Fig. 4—Beam-deflection tube.

Aperture and Phosphor Patterns

Reflection Case

In Fig. 5 is shown a sketch of the mask *ABCE* in relation to the center of deflection *O* of the electron beam. It is obvious that, as the beam scans over the mask in the usual manner, the angle of incidence α is constantly changing. This means that, if d and V_c/V_0 are held constant, S is constantly changing, according to (5). Hence, it would be impossible to excite a single color over the entire screen with a uniformly-spaced array of holes and phosphor dots or straight slits and phosphor lines. If a uniformly-spaced array of apertures and phosphors is to be used, then S must remain constant with changes in α. This could be accomplished by using a curved reflector plate or by modulating the reflector voltage, or by a combination of both. It would appear simpler, however, to keep d constant and, with a fixed value of V_c/V_0, let S vary with α according to (5). A pattern of apertures and phosphors could then be laid out in such a way that it would be possible to excite a single color when the mask was scanned. By changing the reflector voltage, the other colors could then be excited one at a time.

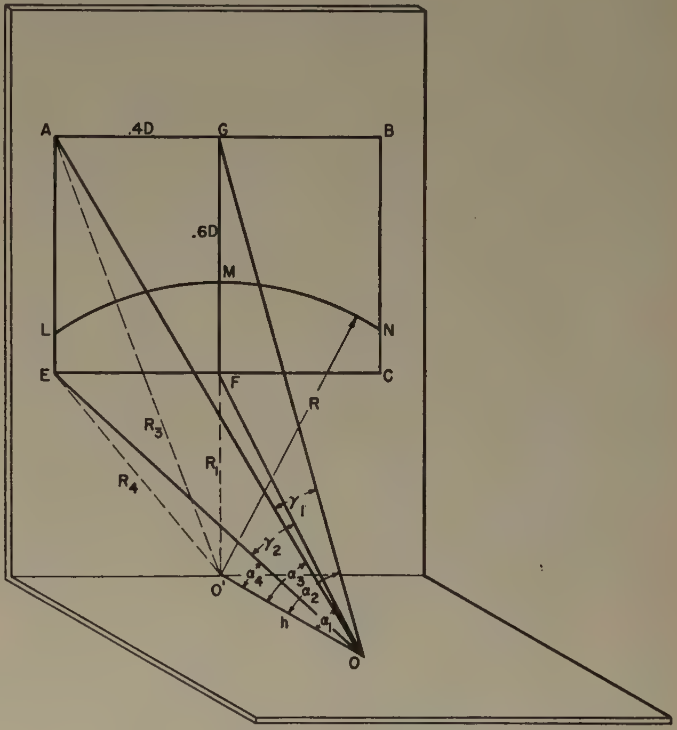

Fig. 5—Geometric relations between the mask *ABCE* and the center of deflection *O*.

It is possible to lay out this pattern in the form of narrow slits that follow contour lines of constant angle of incidence on the screen. In Fig. 5, a perpendicular is drawn from *O*, the center of deflection, to the plane of the screen *O′*. With *O′* as the center, an arc *LMN* of radius *R* is drawn on the screen. When the beam enters the mask anywhere along this arc, it makes a constant angle of incidence α with the mask. It then returns to the mask at a constant distance S above the arc, measured along a radial line that passes through *O′* and the point where the beam enters the mask. The value of S given by (5) may also be written in the form

$$S = 2S_0 \frac{R/h}{(R/h)^2 + 1}, \qquad (34)$$

where $S_0 = 2d_0$, i.e., equal to the displacement of the beam when $\alpha = 45$ degrees. The length of the perpendicular OO' is h. In (34), $\sin 2\alpha$ has been expressed in terms of R and h.

The three phosphors that are to be excited by the electrons that pass through the screen at radius R are laid down in juxtaposition in the form of three stripes, which are arcs of concentric circles with O' as center. It can now be seen how the entire screen may be made up of interlaced slits and tricolor phosphor stripes, all of which are arcs of concentric circles with O' as center.

The fact that displacement of the returning beam will be a maximum at an angle of incidence of 45 degrees, and will become less each side of 45 degrees, means that the slits in the screen will be separated farther near 45 degrees than for greater and smaller angles. It would be desirable to keep the variation in the spacings between the slits as small as possible. The smallest angle of incidence that the beam makes with the screen will be at point F, and the largest angle of incidence will be at points A and B. If the screen is positioned with respect to the center of deflection so that the angle of incidence is 45 degrees somewhere in the center region of the screen, the slits will be spaced closest together at point F and at points A and B. By more exact positioning of the screen, it would be possible to make the slit spacing at F equal to the spacing at A and B. However, from the focusing point of view, it is more desirable to center the mask about a 60-degree angle of incidence and keep to the smallest angle of incidence at F greater than 45 degrees. Under these conditions, the slits will have their greatest spacing at F and their closest spacing at A and B. It is of interest to compute this maximum variation in slit spacings and the various deflection angles involved.

Let it be arbitrarily assumed that the angle of incidence is 45 degrees at F and 60 degrees along a radius whose value is the mean of R_1 and R_3. From the geometry of Fig. 5, the following values are found.

Distance from center of deflection to mask $= h$ $= 0.463D$
Maximum angle of incidence $= \alpha_3$ $= 68$ degrees
Minimum angle of incidence $= \alpha_1$ $= 45$ degrees
Top horizontal deflection angle $= 2\gamma_1$ $= 41\frac{1}{4}$ degrees
Bottom horizontal deflection angle $= 2\gamma_2$ $= 81\frac{1}{2}$ degrees
Vertical deflection angle $= \alpha_2 - \alpha_1 = 21\frac{1}{2}$ degrees
Maximum ratio of slit spacings $= S/S_0$ $= 0.699$.

The value of γ_2 may be reduced and the value of S/S_0 increased at the expense of focusing at the bottom of the mask by making the angle of incidence greater than 45 degrees at F.

The procedure for computing the slits in the mask will now be considered. The radius R_1 of the first slit at point F is determined from the considerations given in the preceding paragraph. The beam, which enters the screen anywhere along this slit, returns to the mask at a distance S_1 from the slit measured along a radial line which passes through O' and the point where the beam enters the mask. The distance S_1 may be computed from (34) by substituting the value of R_1 for R. The radius R_2 of the arc where the returning beam strikes the mask is then given by $R_1 + S_1$. If a slit is placed in the mask at radius R_2, the returning beam from the slit at radius R_1 will pass back through the mask, and a new beam which enters at radius R_2 will return and strike the mask at radius R_3. The value of R_3 may be computed by substituting the value of R_2 for R in (34) and adding the value of S_2 thus obtained to R_2. The general recurrence formula for computing the slit radii is

$$R_{n+1} = R_n + S_n, \tag{35}$$

where

$$S_n = 2S_0 \frac{R_n/h}{(R_n/h)^2 + 1} \cdot \tag{36}$$

With these equations, it is thus possible to lay out a series of slits which are arcs of concentric circles in such a way that the electrons that enter through the nth slit will pass back out through the $(n+1)$th slit.

Additional sets of slits must be interlaced between these so that not less than 25 per cent of the mask area is occupied by slits. The remaining area is provided for the color phosphors. The exact number of slits to be placed between slits of radius R_1 and R_2 will depend on the width of the slits and on the value of S_0.

The positioning of the slits between R_1 and R_2 completely determines the position of the remaining slits on the screen by virtue of (35). It is desirable, therefore, to position the slits between R_1 and R_2 in such a way that the remaining slits will be uniformly spaced without abrupt discontinuities in the spacing. M. H. Hebb of this laboratory has suggested that Newton's interpolation formula may be used for this purpose. It is helpful in understanding this procedure to make use of the graphical interpretation shown in Fig. 6. In Fig. 6(a), S is plotted as a function of R according to (34). The value of R_1 is located on the R-axis, and a vertical line is drawn to the curve. The length of this line is S_1. A dotted line is drawn from point x to the R-axis, making an angle of 45 degrees with it. The intersection of this line with the axis gives the value of R_2. A vertical line from R_2 to the curve gives the value of S_2. This process may be continued until all the values of R are determined from R_1 to R_n. In Fig. 6(b), R is plotted as a function of n. Values of R for integral values of n are obtained from Fig. 6(a). If there are $s-1$ slits between R_1 and R_2, their position can be determined by interpolation of the curve in Fig. 6(b) between R_1 and R_2. This can be done accurately by use of Newton's interpolation formula:

$$R_{1+k} = R_1 + k\Delta_1 + \frac{k(k-1)}{2!}\Delta_2 + \frac{k(k-1)(k-2)}{3!}\Delta_3 + \cdots, \tag{37}$$

where k takes on the successive values $1/s$, $2/s$, $3/s$, $\cdots$ $(s-1)/s$ and the Δ's are the tabular differences determined from the known points on the curve in Fig. 6(b). The recurrence formulas (35) and (36) are now used with the values of R_{1+k} computed from (37) to obtain the positions of the remaining slits.

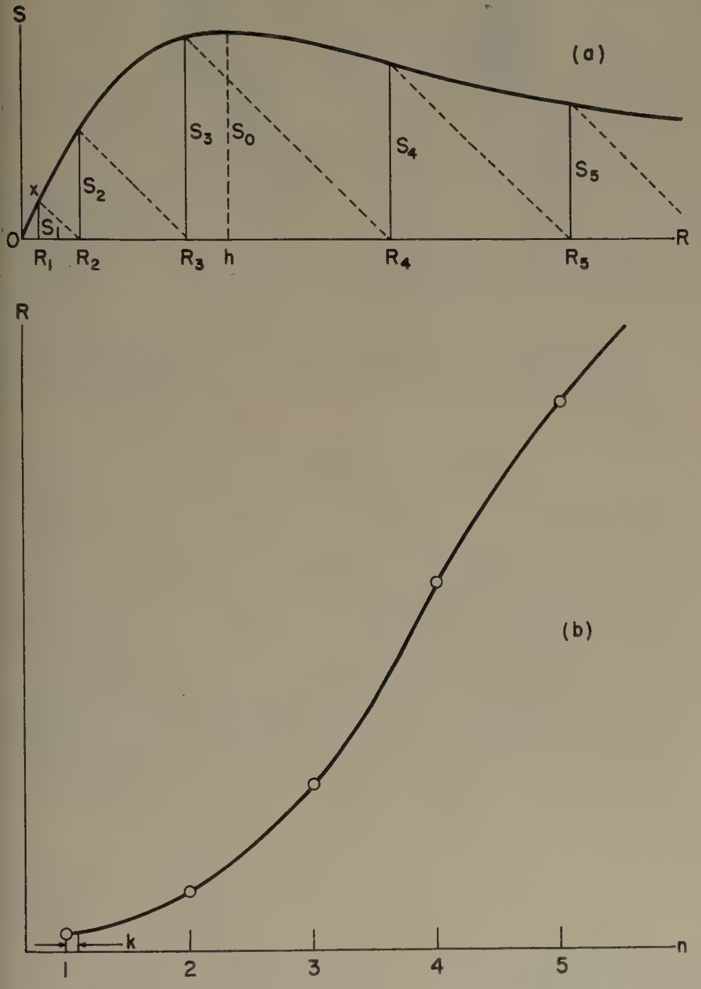

Fig. 6—Procedure for computing location of slits in the mask.

Transmission Case

In this case, the aperture pattern may also be in the form of slits that follow contour lines of constant angle of incidence on the mask. In Fig. 5, with O' as center, an arc LMN of radius R is drawn on the mask. When the beam enters the mask anywhere along this arc, it makes a constant angle of incidence α with the mask. On passing through the mask, it strikes the screen along the arc of a larger circle of radius $R+L$, where L is given by (12). The three phosphors that are to be excited by the electrons that pass through the mask at radius R are deposited on the screen in juxtaposition in the form of three stripes which are arcs of concentric circles. It is thus possible to lay out both the apertures in the mask and the phosphor lines in the form of arcs of concentric circles. In general, for every slit in the mask of radius R_n, there will be a set of three phosphor lines on the screen centered about the radius

$$R_{sn} = R_n + L_n, \tag{38}$$

where L_n, given by (12) is written in the form

$$L_n = \frac{2d}{(1 - V_c/V_0)} \frac{R_n}{(R_n^2 + h^2)} \cdot \{h - [h^2 V_c/V_0 - R_n^2(1 - V_c/V_0)]^{1/2}\}. \tag{39}$$

This type of mask and screen pattern would lend itself to a parallel-plane geometry. The apertures can also be made in the form of parallel horizontal slits by using grid wires. The mask could be curved in one dimension and the screen could be curved in two dimensions. The position of the phosphors could be determined by electron-photography methods.

TOLERANCES AND DESIGN CONSIDERATIONS

Reflection Case

The successful construction and operation of this tube depends almost entirely on being able to maintain certain dimensional tolerances. Voltage variations are of some importance, but to a much smaller degree. The most critical requirement is that of maintaining parallelism between the mask and reflector. The change in displacement of the beam with change in reflector-mask spacing is found by differentiation of (5):

$$\frac{dS}{dd} = \frac{2 \sin 2\alpha}{1 - V_c/V_0} = \frac{S}{d}. \tag{40}$$

It should be noted that dS/dd is greatest for electrons that pass through the screen at an angle of 45 degrees. For these electrons, dS/dd depends only on the ratio of V_c/V_0, decreasing with smaller V_c/V_0. It is impractical, however, to make the reflector voltage less than zero. Under these conditions, $dS/dd=2$. This means, for example, that if the reflector-mask spacing should change one mil, the beam will be displaced two mils. One can thus see the importance of maintaining d constant over the entire screen area. Stated more precisely, the value of d as averaged by the electron in traversing its trajectory should be constant over the entire screen. Small variations in d of a localized nature, such as roughness of the screen surface, will not affect the beam displacement because they do not change the average field appreciably.

It is of interest to plot (5), (9), and (40), as a function of V_c/V_0 for $\alpha=45$ degrees. These curves are shown on the left side of Fig. 7. In the design of a tube, the beam voltage usually will be specified to give a certain brightness level. This leaves two parameters at one's disposal, namely, the reflector voltage and the reflector-mask spacing. From Fig. 7(c), it can be seen that dS/dd is independent of d and is reduced only by re-

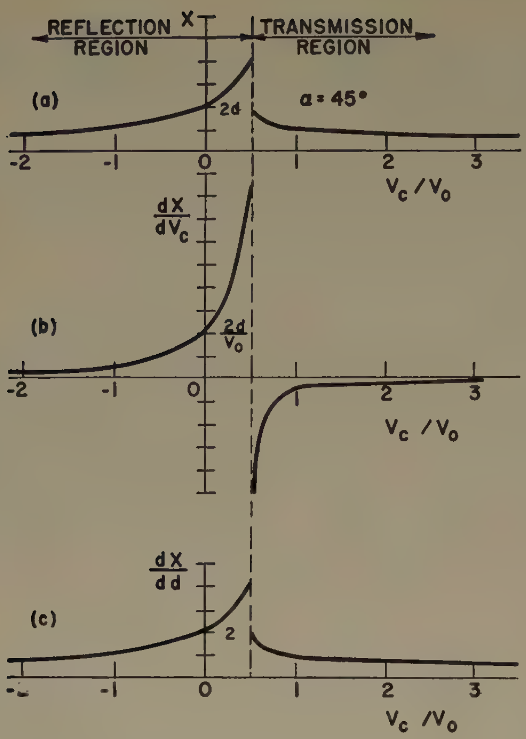

Fig. 7—Curves showing various tube parameters as a function of the ratio of reflector voltage to beam voltage.

ducing V_c. A practical lower limit for V_c is zero. Fig. 7(b) shows that lowering V_c also reduces the deflection sensitivity. However, this may be increased by increasing d. There is a limit as to how far one can go in this direction because, as d is increased, the weak fringing-field region around the edge of the screen extends further into the center portion. It will be observed from Fig. 7(a) that increasing d also increases the beam displacement. A large displacement of the beam leaves a wide band of unused screen area at the bottom of the tube. Small variations in slit-spacing due to stretching or expansion of the screen are proportional to S and, hence, become more serious when the beam displacement is large. There is one advantage in having the beam displacement large, however, besides the reduction in color-switching voltage. Eq. (5) can be written in the form

$$S = \frac{2V_0 \sin 2\alpha}{E}, \tag{41}$$

where E is the electric-field strength between the reflector and screen. It thus appears that the electric field varies inversely as the beam displacement. It is desirable to keep this field low to prevent field emission from the reflector plate and to reduce the electrostatic force that tends to bow the screen. The magnitude of this force is given by

$$F = 4.51 \times 10^{-10} E^2, \tag{42}$$

where F is in grams per square centimeter and E is in volts per centimeter.

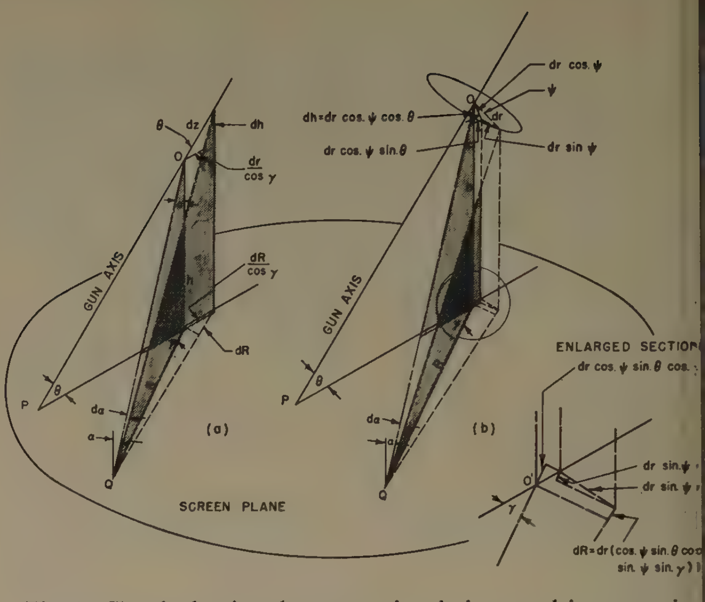

Fig. 8—Sketch showing the geometric relations used in computing the change in beam displacement due to a dislocation of the center of deflection.

It will be helpful to determine how much the beam displacement will change at various points on the screen owing to dislocation of the center of deflection. It is convenient to break this dislocation up into two components, one in the direction of the gun axis and the other, a radial component, at right angles to the gun axis. This first component will change as the deflection yoke is moved along the neck of the tube, and the second component may be changed by applying a traverse magnetic field to the beam before it traverses the deflection yoke.

Since S is a function of both R and h,

$$dS = \left(\frac{\partial S}{\partial R}\right) dR + \left(\frac{\partial S}{\partial h}\right) dh. \tag{43}$$

Partial differentiation of (34) gives

$$\frac{\partial S}{\partial R} = -\frac{2S_0}{h} \frac{(R/h)^2 - 1}{[(R/h)^2 + 1]^2}, \tag{44}$$

and

$$\frac{\partial S}{\partial h} = \frac{2S_0}{h}\left(\frac{R}{h}\right) \frac{(R/h)^2 - 1}{[(R/h)^2 + 1]^2}. \tag{45}$$

From Fig. 8(a), it can be seen that

$$dR = \cos\gamma \cot\theta dh \tag{46}$$

and

$$dh = \sin\theta dz. \tag{47}$$

Substituting these quantities in (43) gives

$$\frac{dS}{dz} = \frac{2S_0}{h} \cos^4\gamma(\tan^2\alpha - 1)\sin\theta(\tan\alpha - \cos\gamma\cot\theta). \tag{48}$$

Similarly from Fig. 8(b),

$$dR = (\cos\psi \sin\theta \cos\gamma - \sin\psi \sin\gamma)dr \tag{49}$$

and

$$dh = -\cos\psi \cos\theta dr. \tag{50}$$

Hence,

$$\frac{dS}{dr} = \frac{2S_0}{h}\cos^4\gamma(\tan^2\alpha - 1) \cdot [\sin\psi\sin\gamma - \cos\psi(\tan\alpha\cos\theta + \cos\gamma\sin\theta)]. \tag{51}$$

In a practical tube design, the trigonometric function in (48) will not exceed about 0.1 for any point on the face of the tube. If d/h is assumed to be about 0.1 and $V_c=0$, then $dS/dz=0.04$. This means that for every 25 mils of motion of the deflection yoke along the neck of the tube, the beam will be displaced 1 mil near the corners of the picture. For similar assumptions, a radial displacement of the center of deflection can produce a beam displacement of about $\frac{1}{6}$ as much. Eqs. (48) and (51) show the advantage in keeping the beam displacement small so far as tolerances in the location of the center of deflection are concerned.

Transmission Case

Plots of (12), (15), and the derivative of (12) with respect to d, are made on the right-hand side of Fig. 7. For the same electrode spacing, it can be seen that the beam displacement and hence the deflection sensitivity and change in displacement with change in electrode spacing are much lower than for the reflection tube.

CONSTRUCTIONAL DETAILS

General Discussion

Only tubes of the reflection-type were constructed; however, both types of tube have some very definite advantages. Since only one electron beam is required, a single gun may be used in a neck of conventional size with standard sweep and focus coils. The procedure for gun alignment is greatly simplified and the probability of cathode failure is also reduced. Most important of all, auxiliary static and dynamic converging equipment is not required because the colors are always in registry. Since the gun arm makes an angle of less than 45 degrees with the screen, the projected length of the tube is decreased to approximately the face-plate diameter, thus permitting the tube to be installed in a shallow cabinet. Both types of tube display equal numbers of red, green, and blue phosphor lines, and the capacitance of the color-switching electrodes are within reasonable limits.

An advantage of the transmission-type tube over the reflection-type tube is efficient use of the electron beam brought about by post-acceleration focusing. The mask for the transmission tube could be made about 80 per cent transparent, while the reflection tube is limited to a mask which is only about 25 per cent transparent because space must be provided on the mask for the three phosphors.

The main advantage of the reflection-type tube is the screen assembly. Since the phosphors are printed on the aperture mask itself, accurate registry between the two is not required in the final tube assembly. The masks are printed before mounting, with the etching and printing all being done from the same masters, thus offering the advantages of mass production. In the reflection tube, the electron beam is reflected back and strikes the screen at positions that are displaced only about one inch from where the beam entered the screen. For this reason, only changes in the dimension per unit length of the screen are important in causing color distortion. In tubes in which the screen and aperture mask are separate, it is the total change in the dimensions of the screen which must be considered. Thus, in large screen sizes the reflection-type tube has a decided advantage.

During the progress of our work on the reflection-type tube, Weimer and Rynn[5] described a similar tube. The tubes made by them contained 7-inch screens and were of an experimental nature to demonstrate the reflection principle. We have made reflection tubes with 16-inch and 24-inch screens and, in doing this, have encountered and solved many problems not included in the preliminary work of Weimer and Rynn. Among these were a method of preventing microphonics in the screen; printing methods that made webbing lines invisible; glass spacer blocks that would withstand high voltage gradients; a method of etching slits in the screen at an angle to reduce beam-current interception; screen alloys that would withstand high-power dissipation; and a high-current electron gun to increase brightness.

Screen Materials and Processing

The screen apertues are in the form of slits which are arcs of concentric circles, laid out according to the calculations given above. The slits are interrupted periodically with webbing lines which hold the screen together and give it strength. These webbing lines are radial lines that emanate from a point in the plane of the screen which is the projection of the center of deflection of the electron beam. The angular separation of the webbing lines is adjusted so that the spacing between them is approximately equal to the spacing between slits at the center of the screen. The line width has been varied from 5 to 11 mils, depending on the screen size. Screens have been made with various numbers of slits ranging from 25 to 40 per inch. The slit widths varied from 5 to 8 mils. The screen thickness has been varied from 3 to 6 mils. A convenient size to work with has been found to be $4\frac{1}{2}$ mils. Since the electrons pass through the screen at an angle, the slits are made oblique so as to intercept a minimum of beam current. This is done by etching the slits in the metal from both sides of the screen at once by a photoengraving process.[6] By offsetting the patterns on the two sides of the screen, it is possible to obtain the oblique slits.

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[5] Weimer and Rynn, loc. cit.

The photographic master used in the photoengraving process was made from full-size rulings. These rulings were made on scribing plates consisting of plate glass coated with a red transparent film. Lines were scribed through this film with Carboloy tools. The rulings were made in the Physical Sciences Development Shops at the University of Chicago, through the courtesy of T. C. O'Donnell. The concentric arcs were ruled to an accuracy of a few ten-thousandths of an inch. The radial webbing lines were ruled on a second plate. Contact prints were made from these plates and combined to give a master positive. The screen metal was coated with a photosensitive resist and exposed to ultraviolet light through the master positive.

A third plate was ruled with continuous arcs of concentric circles, which were midway between the slits. A contact print was made from this plate to give a master for making the zinc lithographic plate used in printing the phosphor lines on the screen. This process is described in detail in the next section. For absolute accuracy, three such plates should be made to give three sets of concentric lines equally spaced between the slits. The error resulting in using a single plate for printing is a crowding together of the lines in some areas. This crowding reaches a maximum at the lower corners of the screen, occurs to a somewhat lesser extent at the upper corners, and does not occur at all along the vertical center line of the screen. It can be easily shown that the spacing between the lines is reduced by a factor equal to the cosine of the angle formed between the vertical center line and a line drawn from the center of curvature of the phosphor lines to the point in question. For a typical screen with the phosphor lines printed on 8 mil centers, the lines will be spaced only 7 mils at the lower corners.

There are many factors that must be considered in selecting a screen material. It must be capable of being rolled into wide thin sheets and must etch readily. In addition, a low coefficient of thermal expansion and a high yield point are desirable. When the screen is stretched on the mounting frame, the amount of elongation that occurs is proportional to the tension, provided the elastic limits of the material are not exceeded. If ΔS is the amount the screen elongates when it is stretched from some minimum tension W_0, below which the screen would not remain flat, to the elastic limit W_m, then

$$\Delta S = (W_m - W_0)/M = \Delta W/M, \tag{52}$$

where M is the Young's modulus of the material.

In practice, the maximum tension that may be put on the screen will be determined by either the strength of the screen (yield point) or the amount of elongation ΔS occurring, depending on the value of the modulus M. If M is low, the screen may be stretched to the point where the elongation will be so great as to cause color misregistry due to change in the spacings between the slits and the corresponding phosphor lines. On the other hand, if M is very large, the screen may rupture on tightening before there is enough elongation to give trouble. In most screen materials, the latter will be the case.

During tube operation, most of the beam power is dissipated at the screen in the form of heat. Nearly all this heat must be lost by radiation. This requires a rise in the screen temperature and results in expansion of the screen. If the temperature rise is ΔT, the expansion per unit length of the screen will be

$$\Delta L = \alpha \Delta T, \tag{53}$$

where α is the coefficient of linear expansion. If ΔL exceeds ΔS, the screen will become loose and buckle. Combining (52) and (53) gives, for the maximum temperature rise,

$$\Delta T = \Delta W/M\alpha. \tag{54}$$

This equation expresses a figure of merit for the screen material. By assuming that the minimum tension is small and that the screen can be stretched to the yield point, it then follows that the power which the screen can dissipate without buckling, and hence the brightness, will vary directly as the yield point divided by the product of Young's modulus and the coefficient of expansion. This, of course, assumes Newton's law of cooling, which holds here because of the small temperature differences involved.

A number of screen materials were tested, including copper, chrome-plated copper, Trodaloy I, Supernickel, cold-rolled steel, and Fernico. Although copper is easily etched, its yield point is too low to be of practical value. Trodaloy I, a copper-cobalt-beryllium alloy containing 97 per cent copper, etches readily, and has about three times the strength of copper. The expansion[7] is 15.8 $\times 10^{-6}$ per degree C., slightly less than copper. This material worked very well in practice. The greatest difficulty was in obtaining sheets of sufficient width. Supernickel, a 70–30 copper-nickel alloy, was comparable to Trodaloy except for strength. The Trodaloy has about twice the strength of Supernickel. The strength of the copper screens could be increased by chrome-plating. However, in order to get strengths comparable to Trodaloy's, the plating thickness had to be increased to the point where it was obstructing the slits in the screen. There was some roughness in the slits caused by plating, which collected phosphors during printing; these phosphors were difficult to remove. Cold-rolled steel had the highest yield point but did not etch well and rusted very easily. Fernico, an iron-nickel-cobalt alloy, had a yield point equal to that of the steel and etched much more readily. Fernico does not rust and has the added advantage of having the very low expansion coefficient of only 4.7×10^{-6} per degree C.

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[6] Screens of this type were made at Buckbee Mears Company, St. Paul, Minn.

[7] These expansion measurements were made by E. E. Burger of this laboratory.

To further increase the power-dissipating ability of the screen, it was coated on the back side with carbon black. It was found that by this procedure the power dissipation could be doubled without buckling the screen.

Phosphor Printing

The phosphors used to obtain the three primary colors were blue calcium magnesium silicate titanium activated, green zinc silicate manganese activated, and red zinc phosphate manganese activated. Although the sulfide phosphors are brighter, the oxide phosphors were found to be more stable and to give color and brightness values that were consistently of good quality.

The phosphors were applied to the metal screen by a process similar to the thermograph process used for imitation engraving. The process consists of printing a set of lines on the screen with a clear vehicle and then dusting them with a phosphor powder. After the lines are completely covered, the excess phosphor is blown off with compressed air. The screen is then placed on the printing press again and a second set of vehicle lines is printed in juxtaposition to the first set. The second set is then dusted with the second color phosphor and the excess blow off. This procedure is then repeated for the third phosphor. The initial printing was done in a hand letter press which was built around a toolmaker's microscope. The copper engraving used in this press was inked with a hand brayer. It was found difficult to print vehicle lines of uniform width with the letter press because the screen and engraving were not absolutely flat. This condition was improved by backing the screen up with soft newsprint, but the condition could never be completely corrected. N. B. Mears, of the Buckbee Mears Company, tried an offset lithographic press in place of the hand letter press with much better results. In this press, the vehicle is picked up by a rubber blanket roll from a zinc etching and transferred to the screen. By proper adjustment of pressures on the roll and screen, it was possible to get a very uniform deposit of vehicle. To obtain the best results, it was found desirable to ink the clean roll twice before making an impression on the screen. The phosphor was dusted on the screen and the excess blown off with compressed air. The screen was baked in air to set the vehicle. This prevented the phosphor from sticking to the blanket when the next impression was made. Fig. 9 shows an example of this type of printing. This method has since been used successfully for printing phosphors for other types of color television tubes.

This method of printing has an advantage in that the size of the phosphor particles is not critical. In the process of dusting, the large grains stick to the vehicle surface in a random way and the smaller ones stick in between, completely covering the surface. This printing process has the further advantage that continuous lines may be printed. This is necessary if the visibility of the webbing lines in the screen is to be kept at a minimum during tube operation.

Attempts to suspend the phosphor as a pigment directly in the vehicle were not successful because the mix lost its ink properties and could not be printed. Spraying a vehicle or phosphor mix onto the screen through a stencil or mask did not give sharp lines because of turbulence in the region of the screen and stencil. A method that did show some promise was that of dusting the phosphor through a stencil in close contact with a screen that had been completely covered with a vehicle. This method gave sharp lines as long as the stencil and screen made intimate contact.

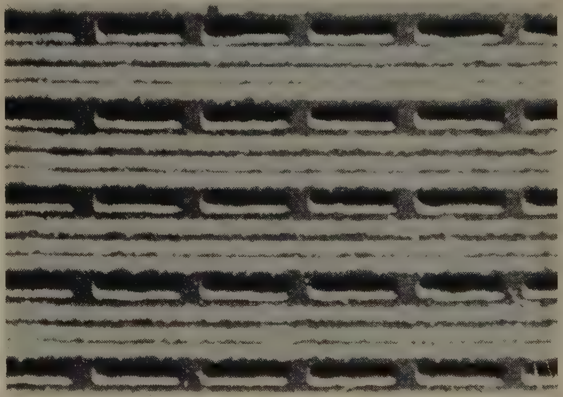

Fig. 9—Photomicrograph of screen showing printed phosphor lines and oblique slits.

Reflectors

The 16- and 24-inch reflectors were made by the Corning Glass Works. Corning ground and polished $\frac{3}{8}$-inch-thick Pyrex for the 16-inch reflectors, and then coated the reflectors with tin oxide. The light transmission was 66 per cent and the resistance was 100 ohms per square. The 24-inch reflectors were ground and polished by Plummer and Kershaw of Philadelphia, Pa., and the Corning Glass Works applied the tin oxide coating. These coatings were all extremely uniform, frequently showing only one interference color across the entire surface.

A silver band around the outer edge of the reflector made electrical contact with the tin oxide. It was found inadvisable to coat silver on the side of the reflector facing the screen because field emission frequently occurred from the surface, giving spurious light on the screen. This band was painted on the reflector before it was sprayed with tin chloride.

A high-resistance tin oxide coating of several thousand ohms per square was placed on the side of the reflector that faces the envelope face plate. This was found necessary to prevent erratic charging of this surface.

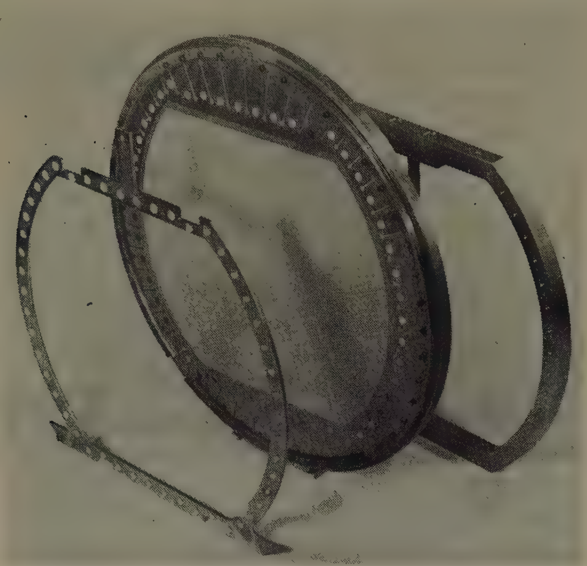

Fig. 10—Screen that has been stretched and is ready for clamping between the mounting frame and ring.

Screen Mounting

The screen is clamped between two thin rings. This assembly is fastened to a large ring by means of screws. The screen is drawn down over the large ring, stretching it tight. The action is similar to that of an embroidery hoop. The stretched screen is then tightly clamped between a heavy mounting frame and a mounting ring by means of screws. The three outer rings used to stretch the screen are then removed and the edge of the screen is trimmed even with the outside of the mounting frame and ring. Fig. 10 shows a screen that has been stretched and is ready for mounting. The three ears on the mounting ring are used to hold the screen assembly in the tube envelope. For the Trodaloy screen and steel frame combination, the screen has a higher thermal expansion than the frame. This caused slight wrinkles to occur at the corners of the screen when it was baked out. These stresses were prevented by slitting the edges of the screen. With the Fernico screens and frames, this was not necessary, and the screen was spot-welded to the frame along the outside edge of the mounting ring.

The tension in the screen was controlled by measuring its resonant frequency. This is done by placing any small light object at the center of the screen and noting the lowest frequency at which it vibrates when the screen is excited acoustically. It was found that the 16-inch Trodaloy screens would rupture at a tension corresponding to 250 cycles. The screens were usually adjusted to a tension corresponding to 210 cycles. The 24-inch Fernico screens were clamped at a tension corresponding to 140 cycles.

Reflector Screen Spacers

To insure proper color registry, the spacing between the screen and reflector must be held to within 0.001 inch. This spacing was made one-half inch for all tubes. This dimension was selected as a compromise among the many factors previously discussed. The reflector is biased at zero volts with respect to the cathode. Thus, the full beam voltage is applied across the $\frac{1}{2}$-inch gap between screen and reflector. The material used for accurately spacing the screen and reflector must withstand this high voltage gradient and should have as low a leakage as practicable.

The best material found for this purpose was a surface-conducting lead silicate glass developed by Blodgett.[8] These blocks measured $\frac{1}{4}$ by $\frac{3}{4}$ by $\frac{1}{2}$ inch, with the high voltage being applied across the $\frac{1}{2}$-inch dimension. The resistance of each block was 3,000 megohms.

The reflector is clamped at each corner between two of these blocks, as shown in Fig. 11. This places eight blocks in parallel between the screen and reflector. At a beam voltage of 20,000 volts, this gives a leakage current of 53 μa. This is not considered excessive, since the beam current is more than twenty times this value.

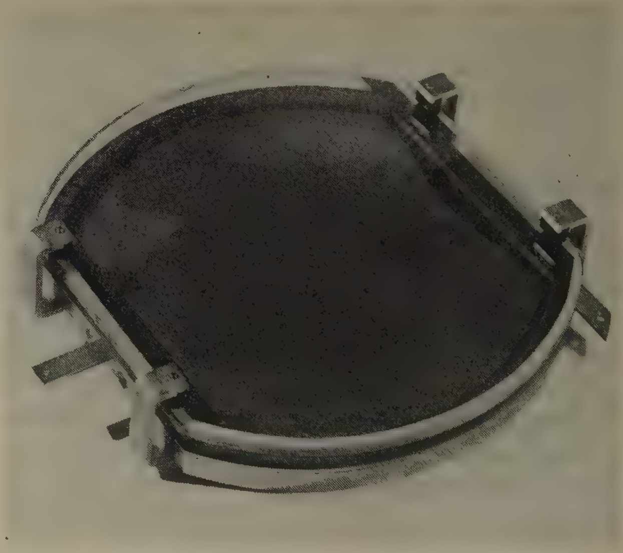

Fig. 11—Screen assembly showing method of supporting reflector-plate between surface-conducting glass blocks.

To insure a uniform potential gradient along the blocks, aquadag is painted along the two edges of the block that are in contact with the reflector and with the metal clamp.

Microphonics

A very close tolerance is required on the reflector-screen dimension to keep the colors in proper registry. An average change of one mil in reflector-screen spacing will change the electron beam displacement by two mils. If the tube is subject to mechanical shock during operation, the screen may be set into vibration, causing beam-displacement modulation and hence color dilution. Since the screen is mounted in vacuum, there is little damping present, and in some cases the vibration may become excessive.

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[8] K. B. Blodgett, "Surface conductivity of lead silicate glass after hydrogen treatment," *Jour. Am. Ceramic Soc.*, vol. 34, pp. 14–27; 1951.

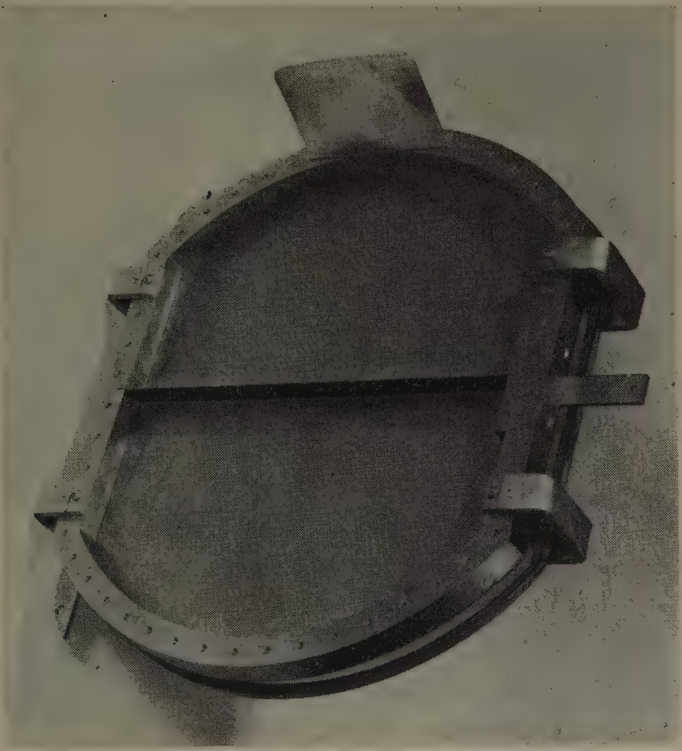

Fig. 12—Back side of screen assembly showing snare used to prevent screen vibration.

To prevent this, a special type of snare was used. It can be seen in Fig. 12. It consists of a thin metal (0.006-inch Trodaloy) strip stretched across the back of the screen, at right angles to it, and spaced about five mils from it. This strip is held under tension by means of spring steel wires mounted in blocks at the ends. The strip and screen are soldered together at their midpoint with a drop of indium. During bake-out, the indium melts and the screen may expand on the frame, becoming buckled and distorted. When the assembly cools down, the screen tightens on the frame, becoming flat again. At 155 degrees C. the indium freezes, holding the screen rigidly in position. The snare is placed directly behind one of the webbing lines to make it inconspicuous during operation of the tube. It can be shown that, if the snare is perpendicular to the screen and is placed along any radial line drawn from the point of projection of the center of beam deflection in the plane of the screen, the shadow will not exceed the thickness of the snare. Actually, since there is no focusing action of the retarding field in a direction at right angles to the snare, the beams spread slightly after passing through the screen on each side of the snare and overlap on striking the screen, thus eliminating the shadow for all practical purposes. Since the webbing lines are all radial lines, as described above, it is only necessary to place the snare over any webbing line to insure proper alignment.

Tube Envelope

The screen assemblies were housed in hemispherical shells. These shells do not necessarily represent the best shape, and certainly do not give the smallest possible volume for the tube envelope. However, since dies were readily available for pressing hemispheres, it was more economical to use this shape in the experimental tubes. The shells are made of the usual 17 per cent chrome-iron alloy used in making the cones of the conventional tubes. In the 16-inch tubes, a glass gun arm and funnel are sealed directly to the hemisphere. The reflector lead seal is also made directly to the side of the hemisphere. The screen assembly is held in position by three mounting posts. These posts are fastened in place by screws welded to the hemisphere.

The neck of the 24-inch tubes is made from a 16-inch cone. The face-plate flange is removed and the cone spun in the form of a cylinder at the end. A glass gun arm and funnel are sealed to the cone. The assembly is coated with aquadag and baked. The gun is then sealed into the neck assembly. The large cylindrical end of this assembly is welded into the 24-inch hemisphere with the aid of a jig attached to the mounting posts. The screen assembly is then mounted and the face-plate assembly welded to the hemisphere as described above. This type of envelope construction has the advantage that all of the glass work may be done on conventional production-line equipment.

Electron Gun

Brightness appears to be a problem in all types of color television tubes at this time. With a single gun tube, the peak current requirements are rather severe, especially if rapid color-switching at megacycle rates is necessary. Under these conditions, sinusoidal waveforms are generally used on the reflector. This means that the beam can only be on about 40 per cent of the time. In practice, only about 25 per cent of the total beam current goes through the slits in the screen and is useful in producing light. Thus, the beam current is reduced to about 10 per cent of its normal value. To get reasonable levels of screen brightness, it is necessary to increase the beam voltage and current. An upper limit of 20 kv was set for the beam voltage, and a special gun shown in Fig. 13 was designed to increase the beam current.

The principle used in this gun is to focus the current from a large extended cathode area through a small, low-voltage, first-anode aperture; that is, to produce a high-perveance gun without provision for modulating the current density by the usual method of varying the electric field at the cathode surface. The current density is modulated by a three-aperture einzel lens following the first anode. The two outer electrodes of this lens operate at first-anode potential. The inner central electrode operates at a negative potential and is used to modulate the beam. This lens acts as a converging lens until the voltage on the central electrode is made sufficiently negative to reduce the axial potential at the center of the lens below zero. At this point, the lens acts as a mirror, reflecting the electron beam.

As the potential of the central electrode is gradually made more and more negative, electrons moving into the lens will not have enough kinetic energy to pass through it, and will be returned. At first, the marginal electrons are returned, but when the central electrode becomes still more negative, the axial electrons are also reflected. In this way, the einzel lens can be used to modulate the intensity of the electron beam. The lens acts as if it were being closed by an iris diaphragm when the central electrode is made more negative.

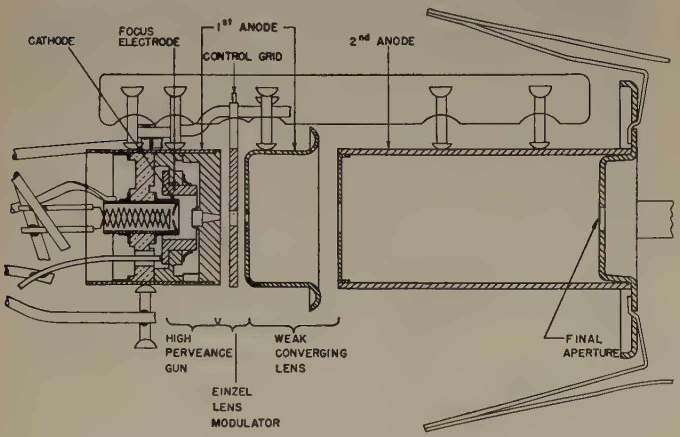

Fig. 13—High-current gun used in reflection-type color tube.

The gun shown in Fig. 13 has a standard $\frac{1}{8}$-inch-diameter cathode. The emitting surface is made spherical, and current from the entire area is focused through a 28-mil hole into the einzel lens. There is a large variety of focusing electrode shapes that may be placed around the cathode to give good focusing action. The central electrode cut-off voltage is determined to a large extent by its thickness and hole diameter. The characteristic curves shown in Fig. 14 are for a gun with a central electrode 30 mils thick and with a hole 60 mils in diameter. The spacing between the central electrode and outer electrodes is 30 mils. Making the central electrode thicker shifts the curves to the right with little change in slope, while making it thinner shifts the curves to the left and also reduces the slope to some extent.

A small opening in the center of the cathode assists alignment and permits the positive ions to pass through without disturbing the cathode emission.

The first aperture on the second anode is for the purpose of increasing the strength of the weak lens action between the first and second anodes. This reduces the divergence of the beam entering the second anode and permits a larger fraction of it to pass through the final aperture. Magnetic focusing and deflection is used.

In some of the tubes, the guns were made with two control grids. One grid was used for the video signal and the other for the 10-mc sampling frequency. The grids were constructed by inserting two adjacent einzel lenses into the electron beam. The maximum beam current for these guns was reduced from five to about three ma.

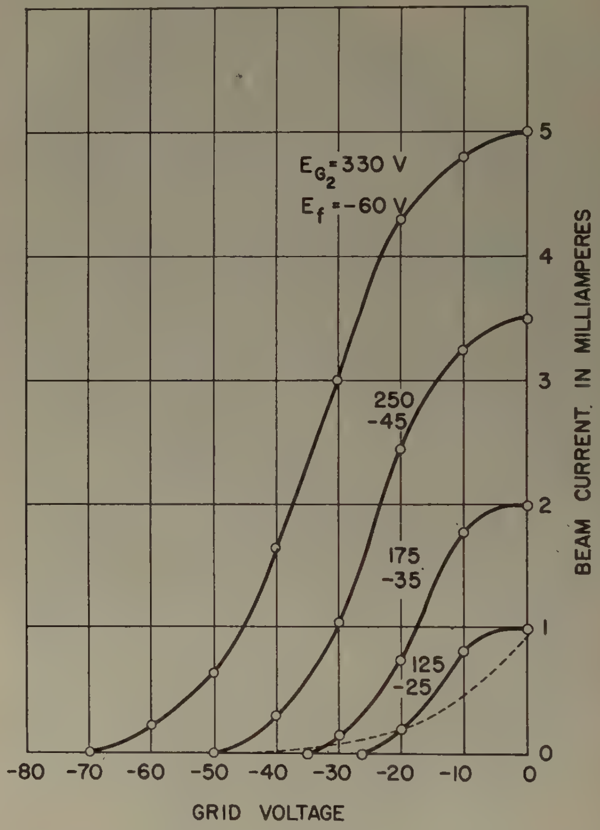

Fig. 14—Characteristic curves for the high-current gun. Dotted curve is for a standard gun used in black-and-white picture tubes.

Results

The sealed-off tubes shown in Fig. 15 were given preliminary tests at the Research Laboratory and were then sent for further evaluation to the Electronics Division at Syracuse, N. Y., where complete color signals are available. In the Research Laboratory, the tubes were tested in a simple setup whereby a black-and-white video signal from a local TV station was applied to the tube. A keystoning correction circuit developed by W. N. Coffey was used with the scanning circuit to give a rectangular raster. By changing the reflector voltage, the picture thus obtained could be made all red, green, or blue. This made it possible to observe brightness, color purity, moiré effects, and resolution.

Nearly all of the 16-inch tubes gave a uniform color field in each of the three primary colors without the aid of correcting fields. For the larger tubes, a beneficial effect was obtained in some cases by placing a permanent magnet near the periphery of the face plate. The color purity was not measured quantitatively, but in general it appeared to be good. Moiré patterns resulting from the beating between the scanning lines and the slits in the screen could be observed at very low beam currents, but at normal picture brightness they were absent. The contrast appeared to be good. However, on close examination, a faint greenish-white background could be observed, owing to scattered electrons.

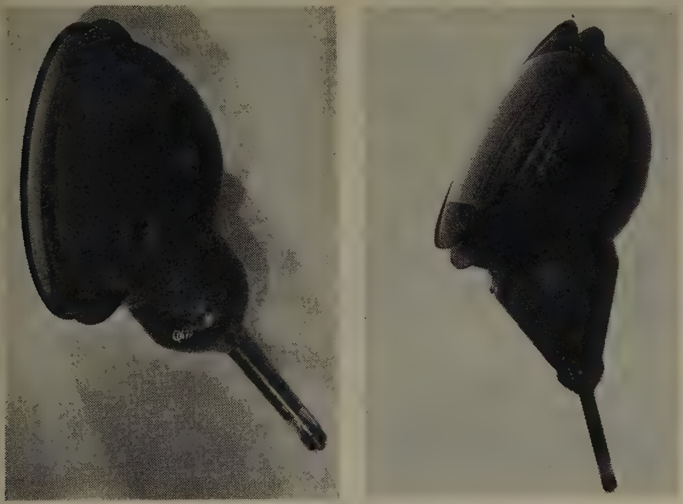

Fig. 15—Completed 16- and 24-inch tubes.

The automatic registry of the three primary colors over all parts of the screen is an outstanding advantage of the reflection-type tube. Besides making the tube simple to adjust and operate, the automatic registry improves the picture resolution. The webbing lines were not visible in the picture at normal viewing distance and did not appear to degrade the resolution. The tube could be operated with average beam currents from $1\frac{1}{2}$ to 3 ma without the 24-inch screens buckling. With the raster just filling the screen, the field brightness was 1 foot-lambert per 100 μa beam current. The characteristids of the 24-inch tube are listed below.

First anode voltage (E_{G_2})	250v
Cut-off voltage	-50v
Beam current, zero bias	1.5 to 3.0 ma
Beam voltage	20,000v
Reflector bias	0 ± 400v
Reflector color-switching voltage	215v rms
Reflector Q (at 3.6 mc)	50
Reflector-screen capacitance	225 $\mu\mu$f.

The tube was tested further with a color-bar generator setup developed by Coffey. Four sets of three color bars were displayed on the screen, using a dot-sequential method in which the colors were switched at a 3.6-mc rate. A 3.6-mc sine wave was applied to the reflector and its third harmonic was applied through a phase-shift network to one of the control grids in the gun for gating the beam. It was found that, by gating the beam in this fashion, the color purity was improved considerably and the brightness maintained, particularly for low-level video signals. The brightness was again found to be 1 foot-lambert per 100 μa of beam current.

APPENDIX

A derivation of the Davisson and Calbrick formula for the focal length of a single aperture in which the electron beam is normal to the plane of the aperture is given in Klemperer's "Electron Optics."[9] The derivation given here follows along similar lines for the more general case in which the incident electron beam makes an angle α with the normal to the plane of the aperture. This derivation was suggested to the author by J. R. Pierce in a private communication.

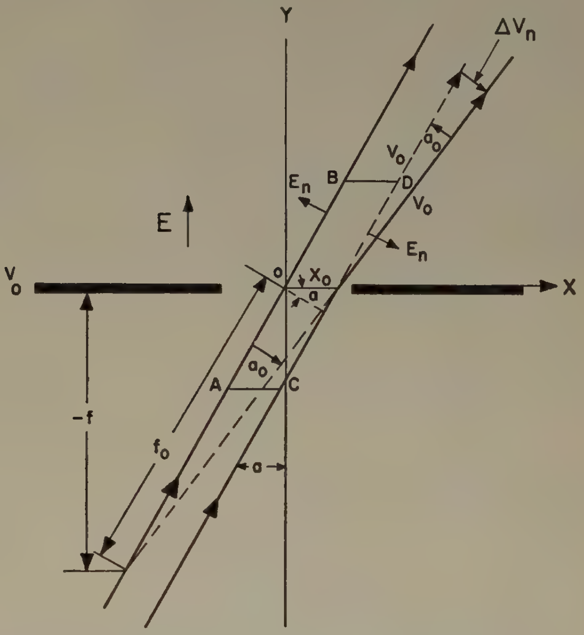

Fig. 16—Derivation of the focal length of a slit.

If the aperture is in the form of a slit, the cylindrical lens field is symmetrical about a plane which passes through the y-axis normal to the x-axis (Fig. 16). Two electron paths will be considered; one through the center of the lens, and the other through the edge of the lens. If it is assumed that the electrons travel with a constant velocity $v_0 = (2eV_0/m)^{1/2}$ through the lens field, then the angle of the path can be changed only through the impulse normal to the path, which produces a velocity change Δv_n normal to the path. Assuming straight-line paths of integration along AB and CD, these changes in velocity are given by

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[9] O. Klemperer, "Electron Optics," University Press, Cambridge, Mass., 2nd ed., p. 62; 1953.

$$\Delta v_{n_1} = -\frac{e}{m}\int(-E_n)dt \quad (\text{along } AB), \tag{55a}$$

and

$$\Delta v_{n_2} = -\frac{e}{m}\int E_n dt \quad (\text{along } CD), \tag{55b}$$

where E_n is the normal component of the lens field along these paths. Substituting $dt = ds/v_0$ and subtracting (55a) from (55b) gives the difference in the normal velocity components for the two paths:

$$\Delta v_n = -\frac{e}{mv_0}\left[\int_C^D E_n ds + \int_A^B E_n ds\right]. \tag{56}$$

These integrals can be evaluated with the help of Gauss's law according to which the surface integral over the normal component of flux density is equal to the inclosed charge. Since there are no components of flux parallel to the slit and there is no charge involved, integration over the path $ABDCA$ gives

$$\int_A^B E_n ds + \int_B^D E_n ds + \int_D^C E_n ds + \int_C^A E_n ds = 0. \tag{57a}$$

Since BD and AC are assumed to be far enough above and below the lens to be in regions of uniform fields E and O respectively, it follows that

$$\int_C^D E_n ds + \int_A^B E_n ds = -Ex_0. \tag{57b}$$

Substituting this in (56) gives

$$\Delta v_n = \frac{e}{m}\frac{Ex_0}{v_0}, \tag{58}$$

from which it follows that

$$\Delta v_x = \frac{e}{m}\frac{Ex_0}{v_0}\cos\alpha \tag{59a}$$

and

$$\Delta v_y = -\frac{e}{m}\frac{Ex_0}{v_0 m}\sin\alpha. \tag{59b}$$

The resultant angular deviation of the electrons is

$$\tan\alpha_0 = \frac{-\Delta v_n}{v_0} = \frac{x_0\cos\alpha}{f_0}, \tag{60}$$

which on substituting (58) and the expression for the velocity gives

$$f_0 = -\frac{2V_0\cos\alpha}{E}$$

or

$$f = -\frac{2V_0\cos^2\alpha}{4E}. \tag{61}$$

It should be noted that (21) and (23) may also be derived from transit-time considerations. Consider the accelerating-field case first. The displacement of the beam is given by

$$L = v_x t, \tag{62}$$

where v_x is the x component of the beam velocity and t is the transit time of the electrons in going from the mask to the screen. In passing through the aperture, the lens field will change the direction of the electrons. This change in direction changes both the x and y components of the initial velocity. A change in the y component of the initial velocity will change the transit time. Hence, from (62),

$$dL = dv_x t + v_x dt. \tag{63}$$

The transit time is given by

$$t = \frac{mv_0}{eE}\left[\cos\alpha - (V_c/V_0 - \sin^2\alpha)^{1/2}\right]. \tag{64}$$

The rate of change of the transit time with respect to the y component of the initial velocity is found to be

$$dt = \frac{m}{eE}\left(1 - \frac{\cos\alpha}{(V_0/V_c - \sin^2\alpha)^{1/2}}\right)dv_y. \tag{65}$$

From (59):

$$dv_x = \frac{eE\cos\alpha}{v_0 m}dx \tag{66}$$

and

$$dv_y = \frac{-eE\sin\alpha}{v_0 m}dx. \tag{67}$$

Substituting (64), (65), (66), (67) and $v_x = v_0\sin\alpha$ in (63) gives

$$dL = \left[\cos^2\alpha - (V_c/V_0 - \sin^2\alpha)^{1/2}\cos\alpha - \sin^2\alpha + \frac{\sin^2\alpha\cos\alpha}{(V_c/V_0 - \sin^2\alpha)^{1/2}}\right]dx. \tag{68}$$

On substituting the condition for focus, $dL = -dx$, in (68):

$$0 = 2\cos\alpha - (V_c/V_0 - \sin^2\alpha)^{1/2} + \frac{\sin^2\alpha}{(V_c/V_0 - \sin^2\alpha)^{1/2}}. \tag{69}$$

Substituting X for $(V_c/V_0 - \sin^2\alpha)^{1/2}$ in (69) and solving for X gives

$$X = 1 + \cos\alpha$$

or

$$V_c/V_0 = 2(1 + \cos\alpha). \tag{70}$$

Similarly, for the retarding field case:

$$S = v_x t,$$

and

$$dS = dv_x t + v_x dt,$$

where

$$t = \frac{2mv_0}{eE} \cos \alpha,$$

$$dt = \frac{2m}{eE} dv_y,$$

$$v_x = v_0 \sin \alpha,$$

$$dv_x = \frac{eE \cos \alpha}{v_0 m} dx,$$

and

$$dv_y = -\frac{eE \cos \alpha}{v_0 m} dx.$$

Hence,

$$dS = 2(\cos^2 \alpha - \sin^2 \alpha) dx;$$

$$dS = 2 \cos 2\alpha dx.$$

Dressler[10] has also derived an expression for obtaining a line-focus with a post-accelerating field. His expression (eq. 2.7), $V_c/V_0 = 3 \cos^2 \alpha + 1$, would appear to be in error because the field has not been integrated over the electron path and the second term in (63) has been neglected. For $\alpha = 0$, his expression gives $V_c = 4V_0$, the same as (23) and (70). For $\alpha = 0$, his path of integration coincides with the electron path and the last term of (63) is zero.

ACKNOWLEDGMENT

The writer wishes to express his appreciation for their advice and encouragement to D. E. Chambers, C. G. Fick, and V. H. Fraenckel, in addition to those already mentioned in the text. Miss J. Gillespie made the numerical calculations for the slit radii of several screens. The active and enthusiastic cooperation of A. T. Wicklund, of the Buckbee Mears Company, in photo-engraving and printing many of the screens, and of J. Sheldon, of the Corning Glass Works, in processing the reflector plates, is gratefully acknowledged. The co-operation of many members of the Research Laboratory's machine shop and glass shop contributed to the construction of the sealed-off tubes. The writer is also very grateful to L. J. Favreau for his valuable help with the experimental work, and in expediting tube construction.

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[10] R. Dressler, "The PDF chromatron—a single or multi-gun tricolor cathode-ray tube," PROC. I.R.E., vol. 41, pp. 851–858; July, 1953.

CORRECTION

G. L. Hall and P. Parzen, authors of the paper, "Measurement of Resonant-Cavity Characteristics," which appeared on pages 1769–1773 of the December, 1953 issue of the PROCEEDINGS OF THE I.R.E., has brought the following correction to the attention of the editors:

The value of R given in the third line of the first paragraph of page 1772 is incorrect. The correct calculation given in more detail is:

$$\left[\int (\Delta\omega/\omega_0)^{1/2} ds \right]^2 = -5.97 \times 10^{-9} \text{ meters}^2.$$

Hence:

$$R/Q_0 = \frac{-1}{2\pi\epsilon_0\omega_0 b^3} \cdot \left[\int (\Delta\omega/\omega_0)^{1/2} ds \right]^2 = 58.1 \text{ ohms},$$

and

$$R = 1.66 \times 10^5 \text{ ohms},$$

where use has been made of the data of Fig. 4 (efficiency of the cavity = 38.8 per cent).

Characteristics of a Transmission Control Viewing Storage Tube with Halftone Display*

M. KNOLL†, SENIOR MEMBER, IRE, H. O. HOOK†, ASSOCIATE, IRE, AND R. P. STONE†, SENIOR MEMBER, IRE

***Summary*—A storage tube is described which maintains a bright visual display, using a charge pattern stored on an insulating layer to grid control the transmitted viewing beam. This insulating layer is a part of an image amplifier system, whose operation and characteristics are discussed. The factors affecting background uniformity and viewing duration have been investigated. Methods of extending the viewing duration by electrically compensating for ion landing have been developed. Erasure, which prepares the surface for the writing of new information, may either be over-all erasure of the entire picture, or selective erasure of only that area where new information is to be written. The present tubes will display a halftone picture of good quality for 30 seconds without compensation, and for over 3 minutes with compensation, after writing has stopped. The writing process can be accomplished in 1/30 second for a complete halftone picture. The resolution is over 500 lines on a 4-inch diameter screen, and there are at least 5 distinguishable halftone steps. As a radar indicator, this tube provides a flickerless display of the complete radar picture, which is bright enough for daylight viewing.**

INTRODUCTION

VIEWING storage tubes are cathode ray tubes in which the visual display of pictures or oscillograms can be maintained for an extended period after writing has ceased. Among different types of such electron devices[1] a grid control tube of the transmission type[2-6] seems to be the most promising today, because of its high brightness and short-erasing time. A transmission-type storage tube, which is capable of reproducing halftones, and permits simultaneous writing and viewing, has been developed.[3,4,6] This paper presents several typical characteristics of the present viewing storage tube.

DESIGN OF TUBE

The basic principles of this type of tube have been discussed in an article in the *RCA Review*.[6] Within the last year a new experimental version of the tube has been built, which is shown in Fig. 1. In addition to refinements which permit better uniformity and greater resolution, this tube differs from the previous one mainly by its cylindrical shape, the inclusion of a separate erasing gun plus the viewing and writing guns, and a larger ratio of screen diameter to over-all bulb diameter.

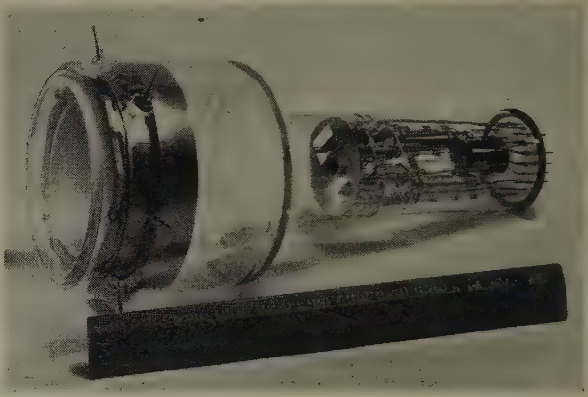

Fig. 1—Photograph of the direct-view storage tube.

Fig. 2 shows the elements of the tube. The writing gun and the erasing gun both carry electrostatic deflection plates and produce focused scanning beams. The viewing gun produces a flooding electron beam. The collimating electron lens insures normal incidence of the viewing and erasing beam electrons at the storage grid.

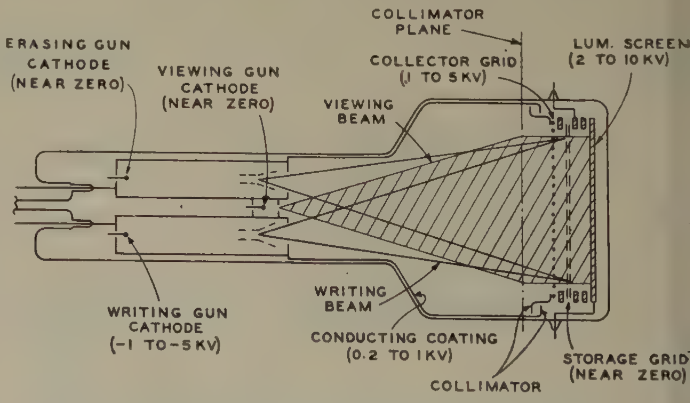
COLLIMATOR PLANE
ERASING GUN CATHODE (NEAR ZERO)
VIEWING GUN CATHODE (NEAR ZERO)
COLLECTOR GRID (.1 TO 5KV)
LUM. SCREEN (2 TO 10KV)
VIEWING BEAM
WRITING BEAM
WRITING GUN CATHODE (-1 TO -5KV)
CONDUCTING COATING (0.2 TO 1KV)
COLLIMATOR
STORAGE GRID (NEAR ZERO)

Fig. 2—Direct-view halftone storage tube (transmission control type).

And finally, at the front of the tube is the image amplifier system, where the picture patterns are written, stored and displayed, and erased.

In this tube the image amplifier system consists of three parallel, closely spaced, electrodes. A collector grid is mounted nearest the guns. The middle electrode is the

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* Decimal classification: R138.31. Original manuscript received by the IRE, May 10, 1954; revised manuscript received, July 6, 1954. The work described in this paper was done at the David Sarnoff Research Center of the RCA Labs. Div., and was supported by a Signal Corps Contract, DA 36-039-sc-15532. A condensed version was presented at the 1954 IRE Convention, New York, N. Y., on March 25, by M. Knoll.

† Radio Corp. of America, RCA Labs. Div., Princeton, N. J.

[1] See, for example, the survey in M. Knoll and B. Kazan, "Storage Tubes and Their Basic Principles," John Wiley & Sons, Inc., New York, N. Y.; 1952.

[2] R. C. Hergenrother and B. C. Gardner, "The recording storage tube," PROC. I.R.E., vol. 38, pp. 740–747; July, 1950.

[3] For details, see M. Knoll, "Electron Lens Raster Systems," and M. Knoll and P. Rudnick, "Electron Lens Raster Viewing Storage Tubes," Washington Symposium on Electron Physics, NBS Circular No. 527; November 5–7, 1951.

[4] Knoll and Kazan, loc. cit., p. 79.

[5] S. T. Smith and H. E. Brown, "Direct viewing memory tube," PROC. I.R.E., vol. 41, pp. 1167–1171; September, 1953.

[6] M. Knoll, P. Rudnick, and H. O. Hook, "Viewing storage tube with halftone display," *RCA Rev.*, p. 492; December, 1953.

storage grid. A luminescent screen on a transparent conducting plate is the third element.

The storage grid is a fine mesh metal screen, in this tube 200 to 400 openings per linear inch. It is covered on the gun side with a thin insulating layer; for example, silica or magnesium fluoride. The charging processes of writing and of erasing take place on this insulating layer, as described subsequently. The charge pattern established in writing modulates the transmission of the flood (viewing) beam over the storage grid surface, and thus a bright image of this charge pattern is produced on the luminescent screen.

We call this target assembly an image amplifier because an increase in power takes place from the written to the viewed picture. Since the writing beam current is 10 microamperes, and the anode current 1 ma, the power of the viewed picture is about 100 times larger than that of the written picture for same beam voltages. Because the stored pattern written in a fraction of a second may be observed for several minutes, the gain in work during such long storage is of the order of 10^6.

Imaging of the Written by the Viewed Picture

Proper imaging of the charge pattern on the storage layer by the picture as observed on the viewing (luminescent) screen is achieved by using each of the storage grid holes as a small electron lens. Each opening focuses that portion of the viewing current passing through it into a separate beam, and to a single spot on the viewing screen. This requires a proper choice of the geometry and the potentials of the image amplifier electrodes. In general, the luminescent screen should be as close as possible to the storage grid to avoid overlapping of the elementary beams, which would produce a loss in useful resolution (see also later section on "Resolution, Halftones, and Brightness"). This image amplifier, therefore, represents in addition, a convenient very short electron optic-imaging system for large images, with the limitation that it gives only unity magnification with the usual parallel grid structures.[7]

Viewing Current Characteristics of the Image Amplifier System

Since the characteristics of the writing, viewing, and erasing guns are more or less conventional, this paper deals primarily with the viewing current characteristics of the image amplifier system.[8]

Characteristics of a Space-Charge Grid Triode

Neglecting for a moment the complications of a double-potential storage grid, we consider first the image amplifier system with a bare control mesh, i.e. a mesh without a storage layer. We may predict that it should have properties similar to Langmuir's[9] space-charge grid triode, which consists of a cathode, one or two accelerator grids, a control grid, and an anode. The characteristics of such a system as measured by Rothe and Kleen[10] are seen in Fig. 3. These authors found that the shape of these characteristics is quite different at high and at low current densities, depending whether electron optic or space-charge conditions prevail near the control grid. In addition, a typical shift of the cut-off voltage takes place within the electron optic region for different maximum anode currents. Equations for characteristics of this kind of accelerating grid triodes exist for the space-charge case, showing the well-known $V^{3/2}$ rise. For the electron optic range, useful equations have been developed only for the case where the field gradients on both sides of the control grid are equal.[10]

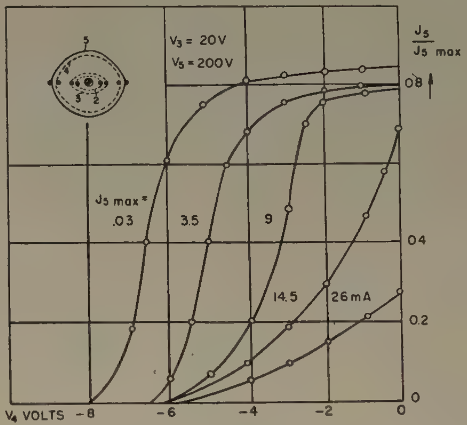

Fig. 3—Characteristic of a space-charge grid tetrode as a function of anode current (Rothe-Kleen).

For the present tube, however, the gradient on the collector grid side has to be low to obtain adequate viewing-beam collimation, and the gradient on the luminescent screen side must be high to get adequate brightness and resolution, and to shift the control characteristic towards more negative V_2-potentials. Because the anode current density of this tube is low (the order of 10 $\mu A/cm^2$), and the field gradient near the control grid holes is high (which presumably causes an "electron optic" current distribution), we expect to find characteristics of the type shown in the left part of Fig. 3.

Method of Measurement

The phosphor in the tube described has a light output proportional to the incident current, within the range of current density used. Therefore, the characteristics may be measured with either brightness or current as the dependent variable. The total current is only an average of the individual characteristics for different sections of the target. By using a photomultiplier which is arranged to "see" only a very small area of the luminescent screen, the characteristic of a particular area may be obtained.

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[7] Knoll, and Knoll and Rudnick, loc. cit.

[8] For the electron optic characteristics, see footnote reference 3.

[9] I. Langmuir, U. S. Patent 1,558,437.

[10] H. Rothe and W. Kleen, "Grundlagen und Kennlinien der Elektronenroehren," Leipzig, Ger., pp. 297–298; 1948. See also, H. A. Pidgeon, "Theory of multi electrode vacuum tubes," *Bell Sys. Tech. Jour.*, vol. 14, pp. 44 and 72; 1935.

Measurements of the characteristics of tubes with bare grids are easily made by a point-by-point method. However, on tubes with storage layers, the viewing duration, due to ion bombardment of the storage layer, is too short for convenient point-by-point measurement. A more rapid method was devised for taking those data by applying a linear sawtooth of voltage from a motor driven potentiometer to the grid, and plotting the brightness or current with an electromagnetic recording oscillograph.[11] A sawtooth period of 10 seconds was found to be a satisfactory compromise between phosphor lag and limited viewing duration.[12]

In recording such characteristics the insulator surface is first charged positive with respect to its backplate by secondary emission. The backplate is then switched negative and the sawtooth voltage started. At the positive excursion of the sawtooth the insulator surface is charged to viewing cathode surface potential which is taken as zero insulator potential. During the second sawtooth a recording is made. By shifting the backplate bias positive in steps a family of characteristics can be recorded.

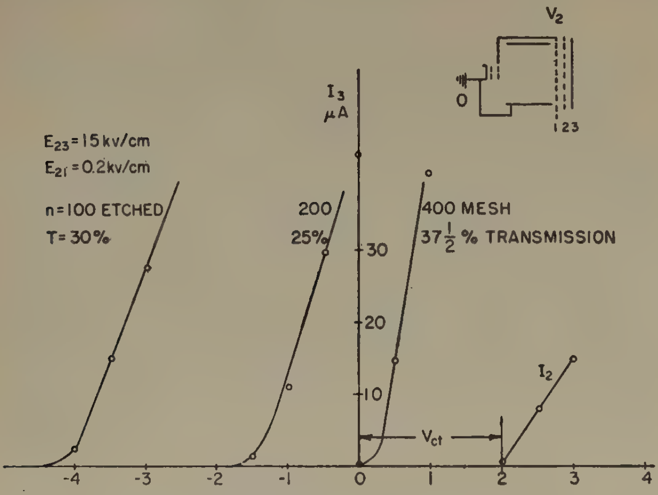


Image amplifier grid potential—V_2—volts (no insulator).

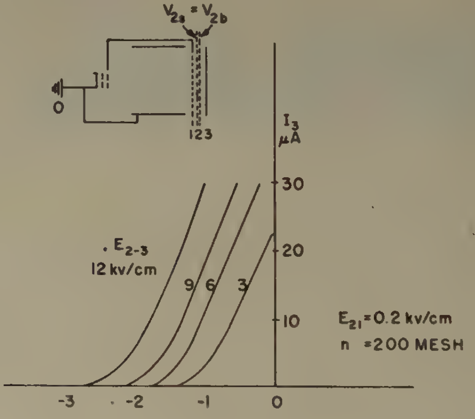
Fig. 4—Viewing current characteristic for bare grids.

Viewing Current Characteristics of Image Amplifier System

The actual image amplifier characteristics for different mesh without storage layers are shown in Fig. 4. As expected, there is close similarity to the low current density curves of Rothe and Kleen. However, there is a substantial shift towards positive grid potentials due to the high amplification factor (1,000–10,000) of the finer mesh, in spite of the 20 times larger gradient.

As can be seen from the grid current curve, storage is possible up to an indicated voltage of +2 volts because the landing current starts only at this high value. This voltage is the sum of contact potential and cathode interface voltage drop. One reason for this high value is that the clean target grid metal is not contaminated by a barium layer as happens frequently in amplifier tubes. Also, since the viewing cathode current is nearly independent of the target grid potential, the cathode interface voltage drop remains high even at low viewing-screen currents.

We may conclude, therefore, that 400-mesh and even finer control grids which may be desirable for high resolution may be suitable for image amplifiers, if the anode gradient is sufficiently high. However, such fine grids may not be necessary from the standpoint of resolution alone. A 250-mesh grid provides adequate resolution for displaying television pictures on a 4-inch viewing screen. Extrapolating to a 12-inch viewing screen, for example, indicates that an 80-mesh grid would suffice for this purpose. Furthermore, the choice of type of grid mesh and of hole spacing may depend on background disturbance, which will be discussed.



Storage grid potential ($V_{2a} = V_{2b}$) volts (recorded).

Fig. 5—Viewing current characteristic for a coated grid with no gradient across the insulator.

Fig. 5 shows the static characteristics recorded for a 200-mesh grid where the storage surface and the metallic grid supporting it are at the same potential. These characteristics were measured as described, making the positive excursion of the sawtooth voltage equal to viewing cathode potential. These curves are plotted only to 30 μamp. so that the region near cutoff may be shown in more detail.

For the same anode gradient, this coated grid characteristic has shifted to the left as compared to the bare grid characteristic. This is due to the method of measurement in which the surface of the insulating layer was set at cathode surface potential by landing viewing-beam electrons. This causes a difference in the zero points taken for the curves by an amount equal to contact potential plus interface voltage drop at cathode.[13]

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[11] Sanborn 2 Channel Oscillograph.

[12] This recording method proved to be more convenient than the oscillographic method used by Hergenrother and Gardner (loc. cit., app. C), which was devised for storage tubes with electrical output.

[13] If the diameter of the grid holes is appreciably decreased by the application of the storage layer, a shift of the corrected curves to the right may occur.

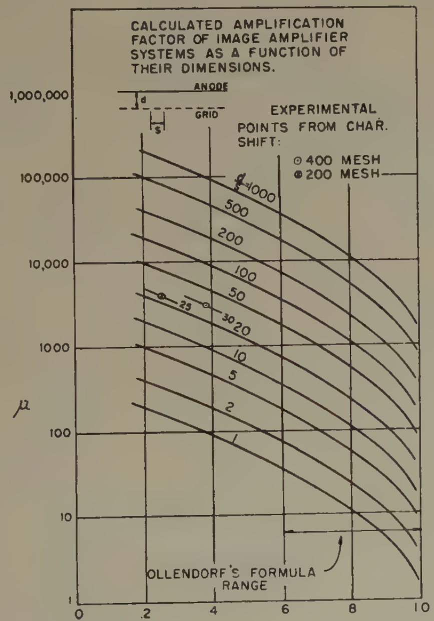
As in the bare-grid case, the characteristics show a typical shift to the left with rising anode gradient. From this shift, the amplification factor, μ, may be determined as $\Delta V_3/\Delta V_2$, and compared with μ values estimated from Ollendorff's formula.[14]

$$\frac{1}{\mu} = \frac{-\ln 2z + z^2}{2a - 2z^2} \tag{1}$$

where $z=\pi\cdot c/s$ ($2c$ wire diameter; s hole spacing) and $a=\pi d/s$ (d anode-grid distance). We introduce in this formula, which was derived for parallel round grid wires, the optical transparency T of the grid as $T=1-A$, where A is the fraction of light absorbing area of the grid. For parallel wires A is equal to the ratio of wire diameter to wire spacing. Therefore, $A=2c/s$ for parallel wires, and

$$z = \frac{\pi}{2}\cdot A = \frac{\pi}{2}(1-T)$$

where T may be measured directly by a photometer, which is more accurate than measuring the dimensions of a fine mesh. For different ratios of anode distance to hole spacing, we obtain a whole family of μ-curves (Fig. 6).

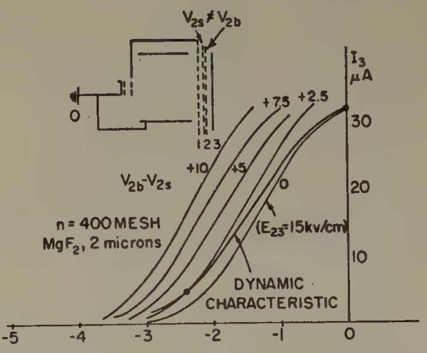


Optical transparency, T.

Fig. 6—Calculated amplification factor as a function of image amplifier dimensions.

Formula (1) assumes relatively high-grid transparencies such as occur in ordinary amplifier tubes. Because it is only valid in a range of $T>0.6$, the question arises whether an extrapolation of these curves would also permit predictions for μ within the low transparency range (10 to 40 per cent) used in these image amplifiers. For a few grids, as shown in Fig. 6, full correspondence was obtained between their measured μ values (obtained from characteristic shift) and the values indicated by an extrapolation of curves calculated from (1). For others, deviations of ±50 per cent have been observed. These deviations are possibly due to differences in the mesh pattern and profile, the effects of which become more important as the hole size becomes small compared to the hole spacing. Such curves, therefore, will probably give accurate μ values only if corrected for a specific mesh pattern and profile. They are, however, very useful for estimates of characteristic shifts.

One might assume that the dynamic characteristic of our image amplifier system is identical with the coated grid characteristic shown in Fig. 5 because only small voltages, of the order of several volts, are necessary to control the anode current. Actually, this is not the case because of the action of the gradient across the storage layer. Since the distance between storage-layer surface and metal backplate is only a few microns, gradients of 10^5 v/cm arise across the storage layer. The action of this gradient near the grid holes shifts the characteristic towards negative V_2-potentials, as if the anode voltage had been raised, for the usual case with backplate positive with respect to the storage-layer surface.



Storage surface potential—V_{2s}—volts (recorded).

Fig. 7—Viewing current characteristic for a coated grid with a gradient across the insulator.

In Fig. 7 a family of such static characteristics is shown in which the potential difference across the storage layer ($V_{2b}-V_{2s}$) is kept constant while the storage surface potential V_{2s} is changed. Comparing Fig. 7 with Fig. 5, it can be seen that 10-volts change in the surface to backplate potential shifts the characteristics by an amount corresponding to an increase of anode voltage V_3 of about 100 per cent.

The dynamic characteristic, therefore, must have a smaller slope than the static characteristic. Its $I_{3\max}$ is the intercept of the static characteristic corresponding to the backplate bias with the $V_{2s}=0$ ordinate. Its lower I_3 values belong to other static characteristics with successively higher gradients.

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[14] See M. Knoll, F. Ollendorff, and R. Rompe, "Gasentladungstabellen," Berlin, Germany, p. 109; 1935, and Rothe and Kleen, loc. cit., p. 85.

For example, the dynamic characteristic for $V_{2b}=0$ is given by the following conditions: $V_{2s}=0$, $V_{2b}-V_{2s}=0$; $V_{2s}=-1$, $V_{2b}-V_{2s}=+1$; $V_{2s}=-2$, $V_{2b}-V_{2s}=+2$; $V_{2s}=-3$, $V_{2b}-V_{2s}=+3$; etc.

The smaller slope of the dynamic characteristic is advantageous with respect to the signal to disturbance ratio. Furthermore, the use of a higher potential difference, $V_{2b}-V_{2s}$, than actually needed for grid control, presents a new possibility of shifting the dynamic characteristic to more negative surface potentials. For a fine mesh grid this might prevent electron landing, and permit using smaller anode gradients. For the grids investigated, the shape of the dynamic characteristics changes only slightly with increasing $V_{2b}-V_{2s}$.

The characteristic of the over-all writing-viewing process is the product of the writing-gun grid-control characteristic and the image-amplifier dynamic characteristic, assuming linear charging of the insulator surface by the writing-beam current. With the writing-gun used in these tubes, the control-grid voltage range is of the order of 40 volts for complete writing in one television frame time.

BACKGROUND DISTURBANCES

In the choice of fine mesh grids, resolution is not the only factor to be considered. Equally important is the background uniformity, which can be measured as the range in storage grid cut-off voltage for the various areas of the screen. This situation is similar to that of a variable mu receiving tube with a luminescent anode, where the cut-off voltages corresponding to areas of different grid pitch could be observed.

The main causes of such background disturbances in image amplifiers are undesired deviations in storage grid dimensions, imperfect collimation, and stray magnetic fields. Storage grid imperfections usually appear as irregular background blotches. Poor collimation usually causes axially symmetric shading. Stray magnetic fields may cause shading or streaks across the tube.

Magnetic disturbances of the low velocity viewing-beam electrons can be reduced sufficiently by demagnetizing the grid frames and by proper magnetic shielding of the bulb. Shadow and moire patterns can be controlled by suitable choice of grid mesh and orientation. An extremely uniform storage grid mesh is needed to achieve good background uniformity.[15]

Other background disturbances occur during the writing process, and will be described. It is notable that in general axially symmetric background disturbances, especially shading, appear much more tolerable to the observer than nonsymmetric ones.

WRITING PROCESS

The present writing process is of the current modulation, nonequilibrium type. It involves two steps: (1) The storage grid surface is shifted to a uniform potential of several volts negative with respect to the viewing cathode potential. This process is the over-all erasing process, which will be described. (2) The storage-grid surface is scanned by the current modulated writing beam. The writing-beam lands on the surface of the storage layer at a voltage between the crossovers of the insulator's secondary emission characteristic, thus causing each element to shift positive by emission of secondary electrons an amount proportional to the instantaneous value of this writing-beam current. The shift is only a few volts toward an equilibrium (collector) potential of the order of 100 volts, so that good linearity of charging is achieved. The useful range of storage grid surface potential is between the negative potential established in step (1) and zero, or viewing beam cathode, potential. If the writing-beam current drives the surface more positive, viewing-beam electrons will land and return the surface to zero.

In the charge pattern which controls the viewing-beam, therefore, high writing-beam currents correspond to less negative storage surface potentials, and low writing-beam currents to more negative storage surface potentials. Thus, the viewing-current pattern is a "positive" of the writing-current pattern, and the tube writes white on a black background. The secondary electrons emitted during writing are collected by the collector grid, the backplate of the storage grid, and the conducting luminescent screen.

In order to limit background disturbances due to unwanted variations in the charge pattern, the charging of the storage layer during writing should be uniform over the target area for a constant writing-beam current. This charging, being a secondary emission phenomenon, is proportional to $(\delta-1)$ and inversely proportional to the capacity of the storage layer. For a homogeneous insulator, uniform capacity implies uniform thickness of the storage layer over the surface. Because of the $(\delta-1)$ factor, the same percentage of variation in secondary emission factor δ produces smaller variations of the charging current at high δ's. Therefore, the use of insulators with a δ large with respect to one, and operation of the writing-beam near the maximum of the secondary emission curve are advantageous. These conditions are also favorable for obtaining a high writing speed. Other factors which may affect the uniformity of the charge pattern are changing of the writing spot size with deflection, nonlinear deflection, and variation of δ with angle of incidence of the writing-beam.

The optimum voltage of the writing-beam is a compromise between electron optic requirements (small focus) and maximum secondary emission. Therefore, the insulator material is chosen to have adequate secondary emission at voltages high enough to produce a sufficiently fine writing-beam.

The writing process can be accomplished in these tubes in 1/30 second for a complete halftone picture. As a single transient oscilloscope, usable traces have been written in a single sweep at the rate of 4 inches per microsecond.

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[15] See M. Knoll, "Background Patterns in Storage Image Amplifier Systems Due to Variations in Amplification Factor and Secondary Emission," N. Y. Nat'l. Conf. Elec. Tube Techs.; Oct. 15, 1953.

USEFUL VIEWING DURATION

For this discussion we define[16] as useful viewing duration T_v the time required for the background brightness to increase to 10 per cent of the highlight brightness. In a halftone picture, this corresponds practically to an increase of the storage layer potential ΔV_{2s} of "dark areas" by an amount of 10 per cent of the useful characteristic range.

For most applications, including radar, viewing durations of 10 to 20 seconds are more than adequate. Viewing durations of 0.1 second are adequate for television applications. Most insulators suitable for storage grids, like silica or magnesium fluoride, will maintain a charge of a few volts for weeks if the tube is not operated. The factor which limits the viewing duration in the normal operation of this tube is not insulator leakage, but the landing of positive ions produced by the viewing-beam, which discharge the storage grid capacitance. Therefore, the useful viewing duration is given by the equation

$$T_v = \frac{\Delta V_{2s} \cdot C_t}{i_+} \text{ [seconds]} \tag{2}$$

where C_t is the storage layer capacity (insulator surface to backplate capacity) and i_+ the positive ion current landing at the whole negative insulator surface, which is assumed to be constant during viewing duration. i_+ can be measured as a function of tube potentials, tube pressure p, and viewing-beam current i_v.

For a particular tube ($\Delta V_{2s} = 0.5$ V, $C_t = 2 \cdot 10^{-7} F$.[17] $i_v = 0.4$ ma, $p = 10^{-7}$ mm Hg, $i_+ = 5 \cdot 10^{-9}\, a$) we find:

$$T_v = \frac{0.5 \cdot 2 \cdot 10^{-7}}{5 \cdot 10^{-9}} = 20 \text{ sec. for halftone pictures.} \tag{3}$$

It can be seen from (2) that long useful viewing durations require low ion current, high-storage layer capacity, and a small slope of the control characteristic. A higher tolerable ΔV_{2s} can be obtained at the expense of resolution by using a coarser grid mesh, which has a longer control characteristic. Also, this tolerable ΔV_{2s} may be an order of magnitude higher for black and white than for halftone pictures. It should be noted that the relatively high-storage layer capacity of 0.2 μF is important for obtaining a long-viewing duration. Thus, at present, viewing durations of much greater than 10 seconds (without some means of compensation for ion landing) have to be paid for by lower writing speed, less brightness, higher vacuum, reduction of resolution, or the abandonment of halftones. The requirement of high vacuum has been made feasible[18] by using a glass bulb which has a low diffusion rate for atmospheric helium,[19] proper processing, and by using effective low pressure-getter processes. For example, the trapping of residual gas molecules by ionization and acceleration towards a negative collector surface may be used. Such ionization may be accomplished by an ionization gauge built into the tube,[20] or by the viewing-beam itself with other carefully outgassed tube elements as collectors.

COMPENSATION AND REGENERATION

For a constant viewing current, the amount of positive ion current reaching the storage surface can be greatly reduced if the image amplifier collector grid is operated as an ion reflector (Fig. 8). However, the remaining ions coming from the spaces between this collector grid and the phosphor screen still land on the negatively-charged storage layer surface or its backplate.

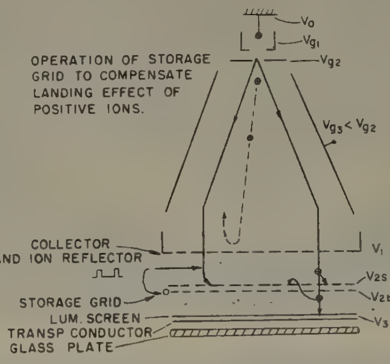

Fig. 8—Positive ion landing, and compensation by electron landing.

Compensation by Backplate Potential Shifting

A rather crude method of partially compensating for the ion landing is to continually shift the dc storage grid backplate potential to more negative values. This potential may be manually adjusted by the operator to approximately compensate for the ion landing, by keeping the dark background constant. Some extension of the viewing duration is possible, but the picture soon deteriorates due to the nonuniformity of the ion landing.

Compensation by Electron Landing

It is possible to compensate for the positive ions by an equal net current of viewing-beam electrons landing below the first crossover of the secondary emission curve. Besides, since the ions are generated by the viewing-beam itself, the ions will land in a pattern which corresponds to the current density of the viewing-beam. Therefore, the electron landing will charge the storage-grid negative in approximately the same pattern that the positive ions have charged it positive. This landing may be achieved by pulsing the storage-grid positive (or viewing-gun cathode negative) at a repetition rate high enough to avoid flicker. Using microsecond pulses at a 60 per second repetition rate extended viewing duration for the same storage tube by a factor of six (from 30 seconds to 3 minutes).

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[16] Viewing duration refers to the over-all performance of the tube, while decay time (see definitions on storage tubes to be published by the I.R.E. Electron Devices Committee) applies only to the storage layer.

[17] $C_t = (kA/11.1\, d)10^{-12}\, F$ where k is the dielectric constant, A the storage grid area, and d the thickness of the storage layer. For $k = 4.4$, $A = 100$ cm^2, and $d = 2 \cdot 10^{-4}$ cm, $C_t = 2 \cdot 10^{-7}\, F$.

[18] For details, see R. P. Stone, "Some Problems in the Design of a Direct View Storage Tube; Techniques for Maintaining High Vacua," N. Y. Nat'l. Conf. Elec. Tube Techs.; Oct. 15, 1953.

[19] See also Lord Rayleigh, *Proc. Roy. Soc.*, London, Eng., vol. 156, p. 350; August, 1936. And D. Alpert and R. S. Burwitz, *Jour. Appl. Phys.*, vol. 25, p. 202; February, 1954.

[20] See G. Krawinkel and T. F. Adams, Fiat Rep. 1021, PB-78273; April, 1947. And D. Alpert, *Jour. Appl. Phys.*, vol. 24, p. 860; July, 1953. Also H. Schwarz, *Zeit. für Phys.*, vol. 122, p. 437; 1944.

Fig. 9 shows the familiar monoscope test pattern. The first picture (#1) was taken immediately after writing. The second (#2) shows the deterioration by positive ions after 30 seconds storage. The third (#3) shows a similar stored halftone picture, with compensation by electron landing, after 3 minutes.

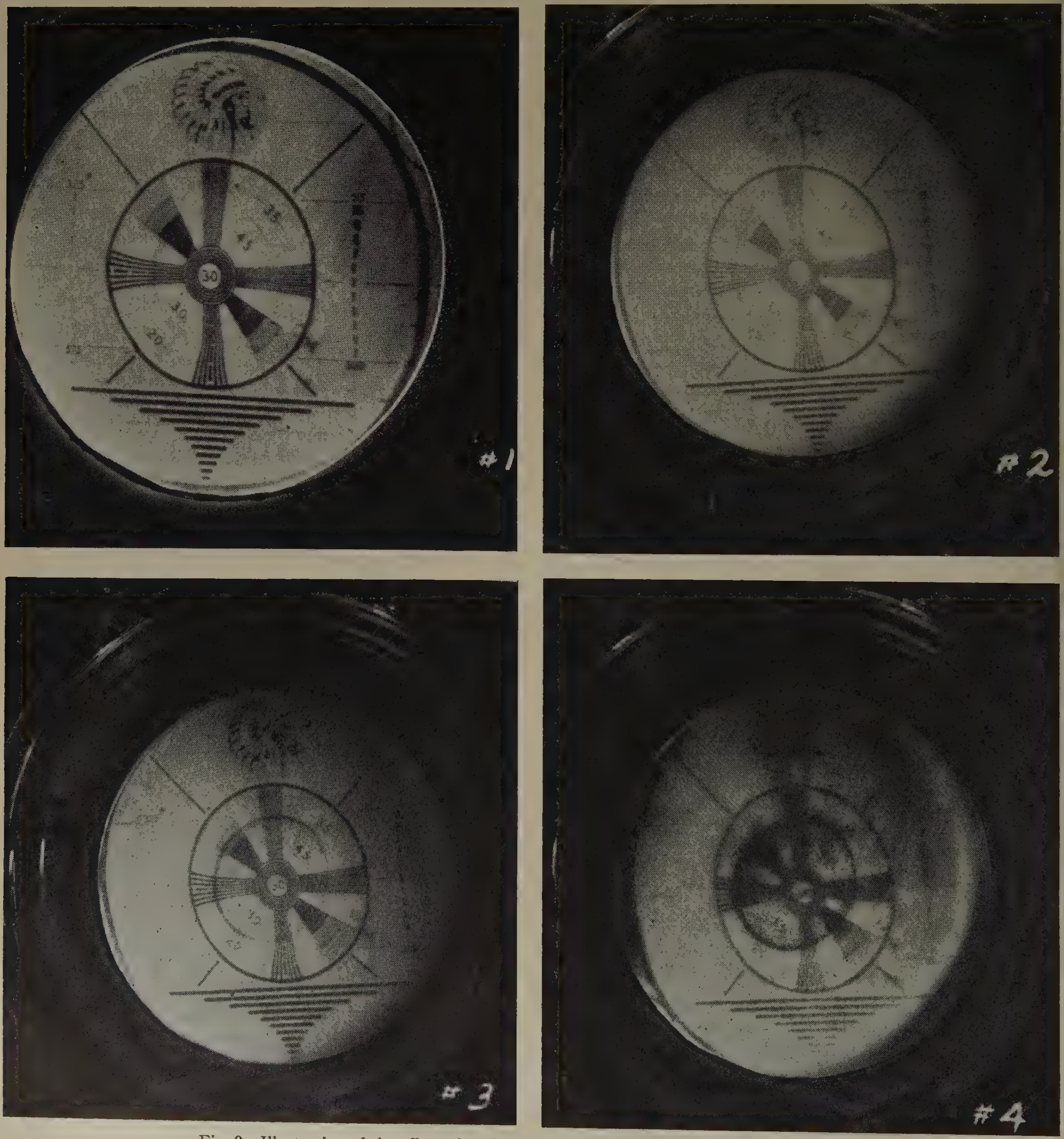

Fig. 9—Illustration of the effect of various methods of extending the viewing duration.

Pulse Holding Method

The fourth picture (#4) refers to another kind of pulse operation which may be called a "pulse-holding" method. This method also serves the purpose of increasing the viewing duration. The potential of the positive pulses applied to the storage grid (or negative pulses to viewing cathode) is chosen slightly higher than the first crossover potential of the secondary emission curve. During the pulse, viewing electrons will land with velocities greater than first crossover at "highlight" storage surface areas, and with velocities less than first crossover at dark storage surface areas. Thus, the more positive areas of the storage layer are driven towards a more positive, and the more negative storage layer potentials towards a more negative potential during the pulse. This pulse-holding method, therefore, is distinct from the static-holding method proposed by Haeff,[21] and the pulse-holding method of Smith and Brown.[22] The latter improves the contrast of a held picture by pulsing the holding (viewing) beam so as to shift the negative ("black") equilibrium potential still more negative. In contrast to Haeff's static method the authors' method may be called a "dynamic" holding method.

Because of the nature of the holding process, this method will not preserve the halftones so effectively as does the compensation method. In the case presented in Fig. 9, it was found that the compensation by electron landing gave better stored pictures up to 3 minutes after which the compensated picture deteriorated. The "pulse-held" picture soon lost much of its halftone quality, but it could be held for hours. Starting with a black-and-white pattern, viewing durations as long as 27 hours have been observed with pulse-holding.

ERASURE PROCESS

Over-all Erasure

The most simple method of erasure consists in switching the storage-grid backplate to a positive potential of several volts after each picture. This potential is below the first crossover potential of the secondary emission curve of the insulator. Viewing-beam electrons will then land and obliterate any charge pattern by charging the storage layer uniformly negative to viewing-cathode potential. If the viewing-screen potential is left on, the screen is illuminated over-all during this erasing process. After switching the storage-grid backplate to viewing-cathode potential, a homogeneous negative potential of several volts remains on the storage surface. This cuts off the viewing beam, and the screen is dark over-all. The storage surface is then ready for the writing process where it will be discharged by an amount proportional to the writing-beam current.

This "over-all erasure" is useful where one stored picture is looked at for a relatively long time. In some cases, however, the flash produced by a long positive-erasing pulse, and the loss of the whole picture instead of only the part which is no longer needed, are serious disadvantages.

Such disadvantages are largely avoided if pulses are applied to the storage-grid backplate during the viewing of the picture, for example, at a repetition rate near television line frequency for a television display. For this purpose, short pulses (of the order of microseconds) are used, and less than one per cent of the picture is erased for each erasing pulse. Such "continuous" over-all erasing was successfully employed for a bright reproduction of television pictures without noticeable "smear."

Over-all pulse-erasing is also practical for radar when erasure over several frames is desirable. This incomplete erasure permits the observation of "tails" on moving objects. When tails are not required, i.e. complete erasure in one frame is desired, over-all erasure may produce objectionable shading at low repetition rates.

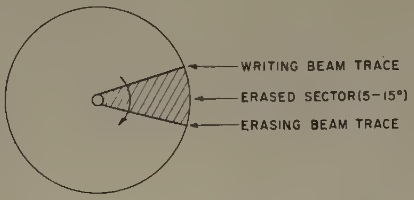

Fig. 10—Bright radar picture with line-by-line erasure.

Line-by-line Erasure

Therefore, a "line-by-line" erasure method was developed which makes use of a separate erasing-beam which may move a short distance in advance of the writing-beam (see Fig. 10). Careful collimation to insure perpendicular landing of the erasing-beam at the storage surface was found to be necessary. Moving objects may be recorded with or without tails, depending on the degree of erasure, which is proportional to the erasing-beam current. In a radar display of the ppi type, the erasing line is visible as a gray trace at the leading edge of the dark erased sector, which is 5 to 15 degrees in width. The writing-beam trace, which moves along the trailing edge of the erased sector, is practically invisible. This is in contrast to present long persistence kinescopes, which write at high brightness, and soon decay. Thus, a flickerless bright display is achieved, which permits daylight viewing of the whole picture except the few per cent of picture area lost in the erased sector.

RESOLUTION, HALFTONES AND BRIGHTNESS

Fig. 9 shows the present quality of the stored pictures. As seen on the face of the tube, the resolution wedge showed a resolution of about 500 lines; in the Indian's head, the individual lines of the 525-line television raster appear. There are also at least 5 distinguishable halftones. Halftones are essential not only for applications such as television but also for detection of radar targets in noise by integration. The maximum contrast ratio of the stored picture (measured with a photomultiplier tube) is at present 20 db (100:1). A highlight brightness of 500-foot lamberts has been measured with 9 kv viewing-screen potential.

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[21] A. V. Haeff, "A memory tube," *Electronics*, vol. 20, p. 80; September, 1947.

[22] Smith and Brown, *loc. cit.*, p. 1171, operation method (d).

In the original picture the focused elementary dots of the 250-mesh storage grid are also visible. This indicates that the image amplifier section is capable of considerably higher resolution than is displayed in the over-all performance of the present tubes, where the writing-beam spot size determines resolution.

Applications

At present, the most promising applications of the viewing-storage tube are airborne, shipborne, and ground-based radar, where its high brightness permits direct viewing in full daylight. Another application is projection television, where this tube is capable of the same brightness as present projection tubes, at much lower anode voltage. Also, the storage properties of the tube make possible television systems conversion. As a single transient storage oscilloscope, usable traces have been obtained with single sweeps of 1 microsecond. Further applications can be seen in closed-circuit facsimile systems, narrow-band television systems, and telemetry.

Conclusion

The characteristics of the image amplifier system in a transmission control viewing-storage tube correspond to those in the "electron optic" region of the well-known space-charge grid triode. Static storage-amplifier characteristics have been obtained by charging the storage-grid surface to various control potentials with the aid of the viewing-beam. It was found for increasingly finer grids that the control characteristic was shifted toward the positive, as may be expected. By using high anode gradients the characteristic is shifted toward the negative. Also, by operating with a gradient across the storage layer, the characteristic is shifted still further to the negative. The effect of these parameters was explored within the useful operating range of the tube.

Dynamic characteristics of the viewing process were obtained indirectly by photoelectric recording of the brightness of an anode area as a function of storage surface potential. Lines of constant backplate potential (the dynamic characteristic) are plotted across the recorded characteristics.

Viewing duration is limited by positive ion landing at the negatively-charged storage surface. The effect is minimized by low residual gas pressure, low viewing current, high storage-layer capacity, and a small slope of the control characteristic. By positive pulsing of backplate and thus compensating for ion landing, viewing duration was increased from 30 seconds to 3 minutes for halftone pictures of 500-line over-all resolution.

Erasing of the entire picture can be effected by the viewing-beam, or individual areas can be erased in a line by line fashion by a separate erasing gun. In each of these modes of operation the erasure can be either complete or partial. Partial erasure produces "tails" on moving targets.

This tube permits, for radar, a new type of flickerless display of the complete picture which is bright enough for daylight viewing.

Acknowledgment

We acknowledge gratefully the support in tube design which was given by Dr. P. Rudnick in the early part of this project. Helpful assistance in measurements was given by Messrs. H. Borkan, W. Carrell, D. H. Fetters, and S. Goodman. Numerous discussions with Dr. D. W. Epstein, B. Kazan, Dr. H. B. Law, Dr. F. H. Nicoll, L. Pensak, Dr. E. G. Ramberg, and Dr. V. K. Zworykin, are also acknowledged.

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Acceptor (in a Semiconductor)—See *Impurity, Acceptor.*

Barrier (in a Semiconductor) (Obsolete)—See *Depletion Layer.*

Base Electrode (of a Transistor)—An *ohmic* or *majority carrier contact* to the *base region.*

Base Region—The interelectrode region of a *transistor* into which *minority carriers* are injected.

Boundary, P-N—A surface in the transition region between *P*-type and *N*-type material at which the *donor* and *acceptor* concentrations are equal.

Carrier—In a semiconductor, a mobile *conduction electron* or *hole.*

Collector (of a Transistor)—An electrode through which a primary flow of carriers leaves the interelectrode region.

Conduction Band—A range of states in the energy spectrum of a solid in which electrons can move freely.

Conductivity Modulation (of a Semiconductor)—The variation of the conductivity of a semiconductor by variation of the charge *carrier* density.

Conductivity, N-type—The conductivity associated with *conduction electrons* in a semiconductor.

Conductivity, P-type—The conductivity associated with *holes* in a semiconductor.

Contact, High Recombination Rate—A semiconductor-semiconductor or metal-semiconductor contact at which thermal equilibrium *carrier* densities are maintained substantially independent of current density.

Contact, Majority Carrier (to a Semiconductor)—An electrical contact across which the ratio of *majority carrier* current to applied voltage is substantially independent of the polarity of the voltage while the ratio of *minority carrier* current to applied voltage is not independent of the polarity of the voltage.

Crystal Pulling—A method of crystal growing in which the developing crystal is gradually withdrawn from a melt.

Depletion Layer (in a Semiconductor)—A region in which the mobile *carrier* charge density is insufficient to neutralize the net fixed charge density of *donors* and *acceptors.*

Diffusion Constant (in a Homogeneous Semiconductor)—The quotient of diffusion current density by the charge *carrier* concentration gradient. It is equal to the product of the *drift mobility* and the average thermal energy per unit charge of *carriers.*

Diffusion Length—In a homogeneous semiconductor, the average distance to which *minority carriers* diffuse between generation and recombination.

Diode, Semiconductor—A two-electrode *semiconductor device* having an asymmetrical voltage-current characteristic.

Donor (in a Semiconductor)—See *Impurity, Donor.*

Doping—Addition of *impurities* to a semiconductor or production of a deviation from stoichiometric composition, to achieve a desired characteristic.

Doping Compensation—Addition of *donor impurities* to a *P-type semiconductor* or of *acceptor impurities* to an *N-type semiconductor.*

Drift Mobility (in a Homogeneous Semiconductor)—The average drift velocity of *carriers* per unit electric field.

Note: In general, the mobilities of electrons and holes are different.

Electrode (of a Semiconductor Device)—An element that performs one or more of the functions of emitting or collecting electrons or *holes*, or of controlling their movements by an electric field.

Electrons, Conduction—The electrons in the *conduction band* of a solid, which are free to move under the influence of an electric field.

Element (of a Semiconductor Device)—Any integral part of the *semiconductor device* that contributes to its operation.

Emitter—See *Emitter, Majority* and *Emitter, Minority.*

Emitter, Majority (of a Transistor)—An *electrode* from which a flow of *majority carriers* enters the interelectrode region.

Emitter, Minority (of a Transistor)—An *electrode* from which a flow of *minority carriers* enters the interelectrode region.

Energy Gap (of a Semiconductor)—The energy range between the bottom of the *conduction band* and the top of the *valence band.*

Extrinsic Properties (of a Semiconductor)—The properties of a semiconductor as modified by *impurities* or *imperfections* within the crystal.

Fermi Level—The value of the electron energy at which the Fermi distribution function has the value one-half.

Forming, Electrical (Applied to Semiconductor Devices)—Process of applying electrical energy to a semiconductor device in order to modify permanently the electrical characteristics.

Generation Rate (in a Semiconductor)—The time rate of creation of electron-hole pairs.

Hall Constant (of an Electrical Conductor)—The constant of proportionality R in the relation

$\mathbf{E}_h = R\,\mathbf{J}\times\mathbf{H}$, where

$\mathbf{E}_h$ = Transverse electric field (Hall field)

$\mathbf{J}$ = Current density

$\mathbf{H}$ = Magnetic field.

Note: The sign of the *majority carrier* can be inferred from the sign of the Hall constant.

Hole—A mobile vacancy in the electronic valence structure of a semiconductor which acts like a positive electronic charge with a positive mass.

Imperfection (of a Crystalline Solid)—Any deviation in structure from that of an ideal crystal.

Note: An ideal crystal is perfectly periodic in structure and contains no foreign atoms.

Impurity, Acceptor (in a Semiconductor)—An impurity which may induce hole conduction.

Impurity (Chemical)—An atom within a crystal which is foreign to the crystal.

Impurity, Donor (*in a Semiconductor*)—An impurity which may induce electronic conduction.

Impurity, Stoichiometric—A crystalline imperfection arising from a deviation from stoichiometric composition.

Intrinsic Properties (*of a Semiconductor*)—The properties of a semiconductor which are characteristic of the pure, ideal crystal.

Intrinsic Temperature Range (*in a Semiconductor*)—The temperature range in which the electrical properties of a semiconductor are essentially not modified by *impurities* or *imperfections* within the crystal.

Junction (*in a Semiconductor Device*)—A region of transition between semiconducting regions of different electrical properties.

Junction, Alloy (*in a Semiconductor*)—A junction formed by alloying one or more *impurities* to a semiconductor crystal.

Junction, Collector (*of a Semiconductor Device*)—A junction normally biased in the high-resistance direction, the current through which can be controlled by the introduction of *minority carriers*.

Junction, Emitter (*of a Semiconductor Device*)—A junction normally biased in the low-resistance direction to inject *minority carriers* into an interelectrode region.

Junction, Fused (*in a Semiconductor*)—A junction formed by recrystallization on a base crystal from a liquid phase of one or more components and the semiconductor.

Junction, N-N (*in a Semiconductor*)—A region of transition between two regions having different properties in *N*-type semiconducting material.

Junction, P-N (*in a Semiconductor*)—A region of transition between *P*- and *N*-type semiconducting material.

Junction, P-P (*in a Semiconductor*)—A region of transition between two regions having different properties in *P*-type semiconducting material.

Junction (*Semiconductor*), *Diffused*—A junction which has been formed by the diffusion of an *impurity* within a semiconductor crystal.

Junction (*Semiconductor*), *Doped*—A junction produced by the addition of an *impurity* to the melt during crystal growth.

Junction (*Semiconductor*), *Grown*—A junction produced during growth of a crystal from a melt.

Junction (*Semiconductor*), *Rate-grown*—A *grown junction* produced by varying the rate of crystal growth.

Lifetime, Volume—The average time interval between the generation and recombination of *minority carriers* in a homogeneous semiconductor.

Majority Carrier (*in a Semiconductor*)—The type of *carrier* constituting more than half of the total number of *carriers*.

Minority Carrier (*in a Semiconductor*)—The type of *carrier* constituting less than half of the total number of *carriers*.

Mobility—See *Drift Mobility*.

Mobility, Hall (*of an Electrical Conductor*)—The quantity μ_H in the relation $\mu_H = R\sigma$, where $R =$ *Hall constant* and $\sigma =$ conductivity.

Ohmic Contact—A contact between two materials, possessing the property that the potential difference across it is proportional to the current passing through.

Photovaristor—A varistor in which the current-voltage relation may be modified by illumination, e.g., cadmium sulphide or lead telluride.

Point Contact—Pressure contact between a semiconductor body and a metallic point.

Primary Flow (*of Carriers*)—A current flow which is responsible for the major properties of the device.

Recombination Rate, Surface—The time rate at which free electrons and *holes* recombine at the surface of a semiconductor.

Recombination Rate, Volume—The time rate at which free electrons and *holes* recombine within the volume of a semiconductor.

Recombination Velocity (*on a Semiconductor Surface*)—The quotient of the normal component of the electron (*hole*) current density at the surface by the excess electron (*hole*) charge density at the surface.

Semiconductor—An electronic conductor, with resistivity in the range between metals and insulators, in which the electrical charge *carrier* concentration increases with increasing temperature over some temperature range. Certain semiconductors possess two types of *carriers*, namely, negative electrons and positive *holes*.

Semiconductor, Compensated—A semiconductor in which one type of *impurity* or *imperfection* (e.g., *donor*) partially cancels the electrical effects of the other type of *impurity* or *imperfection* (e.g., *acceptor*).

Semiconductor Device—An electron device in which the characteristic distinguishing electronic conduction takes place within a semiconductor.

Semiconductor Device, Multiple Unit—A semiconductor device having two or more sets of electrodes associated with independent *carrier* streams.

Note: It is implied that the device has two or more output functions which are independently derived from separate inputs, e.g., a duo-triode transistor.

Semiconductor Device, Single Unit—A semiconductor device having one set of electrodes associated with a single *carrier* stream.

Note: It is implied that the device has a single output function related to a single input.

Semiconductor, Extrinsic—A semiconductor with electrical properties dependent upon *impurities*.

Semiconductor, Intrinsic—A semiconductor whose electrical properties are essentially characteristic of the pure, ideal crystal.

Semiconductor, N-type—An *extrinsic semiconductor* in which the *conduction electron* density exceeds the *hole* density.

Note: It is implied that the net ionized *impurity* concentration is *donor* type.

Semiconductor, P-type—An *extrinsic semiconductor* in which the *hole* density exceeds the *conduction electron* density.

Note: It is implied that the net ionized *impurity* concentration is *acceptor* type.

Space Charge Region (*Pertaining to Semiconductor*)—A region in which the net charge density is significantly different from zero. See also *Depletion Layer*.

Thermistor—An electron device which makes use of the change of resistivity of a semiconductor with change in temperature.

Transistor—An active *semiconductor device* with three or more electrodes.

Transistor, Conductivity Modulation—A *transistor* in which the active properties are derived from *minority carrier* modulation of the bulk resistivity of a semiconductor.

Transistor, Filamentary—A *conductivity modulation transistor* with a length much greater than its transverse dimensions.

Transistor, Junction—A *transistor* having a *base electrode* and two or more *junction* electrodes.

Transistor, Point-contact—A *transistor* having a *ba* *electrode* and two or more *point-contact* electrodes.

Transistor, Point-junction—A *transistor* having a *ba* *electrode* and both *point-contact* and *junction* electrode

Transistor, Unipolar—A *transistor* which utiliz charge *carriers* of only one polarity.

Transition Region—The region, between two hom geneous semiconductor regions, in which the *impuri* concentration changes.

Valence Band—The range of energy states in th spectrum of a solid crystal in which lie the energies the valence electrons which bind the crystal together.

Varistor—A two-electrode *semiconductor device* havin a voltage-dependent nonlinear resistance.

Zone Leveling (*Pertaining to Semiconductor Proces ing*)—The passage of one or more molten zones along semiconductor body for the purpose of uniformly di tributing *impurities* throughout the material.

Zone Purification (*Pertaining to Semiconductor Pro essing*)—The passage of one or more molten zones alon a semiconductor for the purpose of reducing the *im purity* concentration of part of the ingot.

Measurement of Circuit Impedance of Periodically Loaded Structures by Frequency Perturbation*

E. J. NALOS†, SENIOR MEMBER, IRE

***Summary*—A method is described for determining the circuit impedance of a slow-wave structure from experimental measurements of the perturbation in the resonant frequencies caused by passing small beads through the interaction region. From these measurements the gain parameter *C* of such a structure, when used as a traveling-wave tube, can be determined. The results agree with measured experimental gains as well as with theoretical calculations. The method is limited to periodic structures which can be short-circuited and hence made resonant. The principles involved are not new, but their adaptation to traveling-wave tubes is not generally known.**

INTRODUCTION

IN ORDER TO evaluate the performance of a given traveling-wave tube, one must first determine the impedance of the propagating structure, since the gain in db is roughly proportional to the cube root of the impedance (see following section). For some simple structures in which the fields are known, this quantit can be calculated, but for many useful slow-wav structures in which the fields are not known, the circui impedance must be determined experimentally. Tw methods are currently employed to determine the im pedance, and the purpose of this paper is to describe third method useful for periodic structures which ca be made resonant by placing short circuits at suitabl symmetry planes.

One method[1] depends on low-level interaction of tw of the three propagating waves, whereby at some lo beam current the waves all add to zero for a know value of *CN*.

Another method consists of measuring gain as a func tion of beam current, and deducing the gain paramete *C* from the slope of this curve.

A disadvantage of both these methods is the fact tha an electron beam must be made to pass through th structure. Perturbation techniques, on the other hanc

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† Stanford University, Stanford, Calif.

[1] R. Kompfner, "On the operation of the traveling-wave tube a low level," *Jour. Brit. I.R.E.*, vol. 10, p. 283; August/Septembe 1950.

offer possibilities of making impedance calculations from measurements on sample test cavities without the use of an electron beam.

One way of applying this technique consists of measuring the change in phase velocity (or guide wavelength) at a given frequency due to the insertion of a perturbing rod along the interaction region. Some exploratory work in this direction is being done at Stanford University at the suggestion of Prof. E. T. Jaynes.

Perhaps an easier way of applying perturbation techniques consists of measurements on a resonant section of the structure several periods long, from which the impedance can be deduced. Briefly, the method consists of:

1. Measurement of the resonant frequencies of such a structure, from which the phase and group velocity can be obtained.
2. Measurement of the relative fields in the interaction region by measuring the change in resonant wavelength as a function of bead position.
3. Measurement of the absolute value of the E-field at the symmetry plane for a given energy stored in the cavity by measuring the change in the resonant wavelength as a small metallic perturbation is removed from the symmetry plane.

Proper interpretation of these measurements leads to the value of the circuit impedance.

DESCRIPTION OF METHOD

From the simple theory of traveling-wave (TW) tubes, it is known that a beam interacting with a slow wave gives rise to three waves propagating along the structure. Of these, only one increases with distance. It is this wave which is of greatest interest, since it eventually dominates the others. Its amplitude increases exponentially with distance as $e^{2\pi NCx}$, where

N = length of the tube in guide wavelengths,
x = a factor (tabulated by Pierce[2]) whose amplitude depends on beam velocity, circuit loss, and space charge,
C = gain parameter (as given by Pierce[2]), defined by

$$C^3 = 1/4\,\frac{\text{Circuit impedance}}{\text{Beam impedance}} = \frac{I_0K}{4V_0} \qquad (1)$$

$$K = \frac{E_m{}^2}{2\beta_m{}^2P} = \frac{E_m{}^2}{2\beta_m{}^2wv_g} = \text{circuit impedance}, \qquad (2)$$

E_m = effective field seen by beam for the mth space harmonic,
$\beta_m = \omega/v_p$ = phase constant,
v_p, v_g = phase and group velocities,
w = energy stored per unit length,
$P = wv_g$ = power flow,
V_0 = beam voltage,
I_0 = beam current.

The gain of the structure in db is obtained by taking the logarithm,

$$G = A + BCN \qquad (3)$$

where A is a factor accounting for the fact that only a portion of the applied wave sets up the growing wave, and $B = (20)\ 2\pi \log_{10}e = 54.5x$.

The gain in db per guide wavelength is thus BC. The factor B is known from theory, its accuracy depending on how closely operating conditions approach those assumed by theory; the gain parameter C is determined from the knowledge of the rf impedance of the slow-wave circuit.

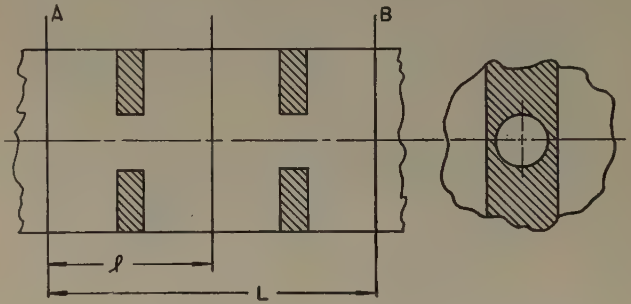

Fig. 1—Typical uniformly loaded slow-wave structure.

Phase and Group Velocity

Consider a section of a uniformly loaded slow-wave structure such as shown in Fig. 1. Form a multiple cavity, comprising n periodic sections, by placing short circuits at the symmetry planes A and B. Such a cavity will have $(n+1)$ resonances whenever the total phase shift is $m\pi (m = 0, 1, \cdots n)$. The Brillouin diagram (frequency versus phase shift per section) could be constructed if one knew the phase shift corresponding to each of the resonances. This phase shift is obtained unambiguously from the number of nodes in the standing-wave pattern by using Slater's perturbation method[3] to determine the axial variation of the field. This is obtained by moving a perturbing object along the direction of propagation and measuring the change in resonant wavelength $\Delta\lambda$. Slater's perturbation formula, rewritten in MKS units, states that

$$\frac{\Delta\lambda}{\lambda} = -\frac{1}{2}\,\frac{\int_{\Delta\tau}(\mu H^2 - \epsilon E^2)dV}{\int_V(\mu H^2 + \epsilon E^2)dV} \qquad (4)$$

where $\Delta\tau$ refers to the perturbing volume and V refers to the cavity volume.

The modes of interest in TW tubes are TM modes which have strong E fields and zero H fields on the axis, and for small perturbing volumes (substantially constant fields over the perturbation), at a given resonant wavelength λ, the relative E field is, from (4)

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[2] J. R. Pierce, "Traveling-Wave Tubes," D. Van Nostrand Company, Inc., New York, N. Y., chap. II; 1950.

[3] J. C. Slater, "Microwave Electronics," D. Van Nostrand Company Inc., New York, N. Y., pp. 80–83; 1950.

$$E_{rel} \sim \sqrt{\Delta\lambda} \tag{5}$$

since the denominator (energy stored in the cavity) is a constant. It is thus possible to obtain a plot of the relative E field along the axis as a function of distance.

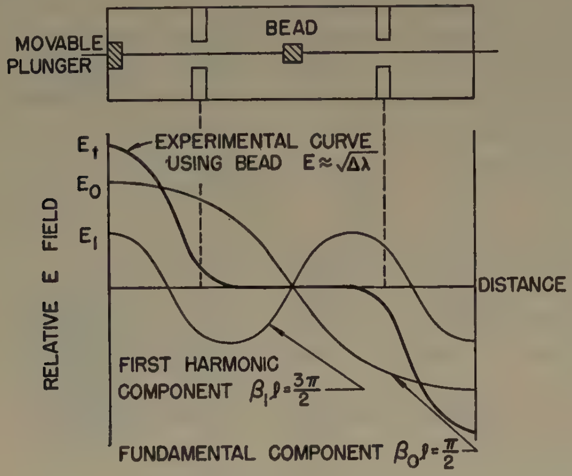

Fig. 2—Field distribution on the axis of a 2-cavity section at the center of the passband.

Such a curve is shown in Fig. 2 for one of the resonant frequencies of a multiple cavity two periods long. The usefulness of this E-field plot is two-fold: from the number of nodes of this curve, the phase shift per section can be determined (in this case one node is observed, i.e., total phase shift of π for 2 sections, or $\pi/2$ per section); and moreover, as described below, this curve is used to determine the effective field as seen by the electron beam. After identification of the phase shift at each resonance, a Brillouin diagram may be plotted. A typical curve is given in Fig. 3 showing the three observed resonant frequencies in a structure two periods long. Two of these frequencies correspond to the cutoff frequencies of the structure; and the third, near the center of the passband, corresponds to a phase shift of $\pi/2$ radians per section. With more sections, a larger number of resonances (i.e., points on this curve) is observed, giving greater detail of the shape of the passband. This technique has been used before in the design of linear accelerators.[4] The usefulness of the Brillouin diagram results from the fact that both the phase and group velocity, required to calculate the impedance, are directly obtainable as a function of frequency from the relations

$$\begin{aligned} v_p &= \frac{\omega}{\beta} \\ v_g &= \frac{d\omega}{d\beta}. \end{aligned} \tag{6}$$

For the case shown in Fig. 3, note that the group velocity (slope of the curve) is negative from $\beta l = 0$ to π, whereas the phase velocity is positive. Such a wave is known as a backward wave. A forward wave in the region of $\beta l = 0$ to π would have been obtained had the side holes been closed off in the structure shown in Fig. 1.

Effective Field Seen by Beam

The remaining quantity required to calculate the circuit impedance is the effective field seen by a beam in its travel through the structure. It is well known that for the unloaded waveguide the field on the axis is sinusoidal with distance and travels with a fixed phase velocity greater than the velocity of light. The loading obstacles not only slow the wave down (to synchronize with a beam of finite voltage) but cause the field on the axis to become nonsinusoidal as shown in Fig. 2 for the case of $\beta l = \pi/2$. If this field is Fourier-analyzed, it is found to contain components each having a different phase shift per section ($\beta_m l = \pi/2, 3\pi/2, 5\pi/2 \cdots$). This is also seen from the Brillouin diagram (Fig. 3) by consideration of a straight line at the resonant frequency f_1, which intersects all the branches of the curve at the above phase shifts per section. The fundamental component, which in this case is a backward wave, travels at a phase velocity corresponding to $\beta_0 l = \pi/2$; the next component, known as the first space harmonic, travels with $\frac{1}{3}$ of this velocity (corresponding to $\beta_1 l = 3\pi/2$); the next, again a backward wave, travels with $\frac{1}{5}$ of this velocity, etc. All these components must be present to make up the total field in the tube at this frequency. The beam may be synchronized with any component whose velocity is less than that of light by a simple adjustment of the beam voltage. The remaining components travel at widely different velocities and their effect averages out to zero. The velocities of each component can be determined directly from the Brillouin diagram, but their amplitudes must be obtained from the Fourier analysis of Fig. 2. The effective electric field as seen by the electron is simply the peak value, E_m, of the particular space harmonic to which the beam is synchronized, since it effectively sees a constant field of this amplitude. The ratio of E_m, to E_t, the total field at the

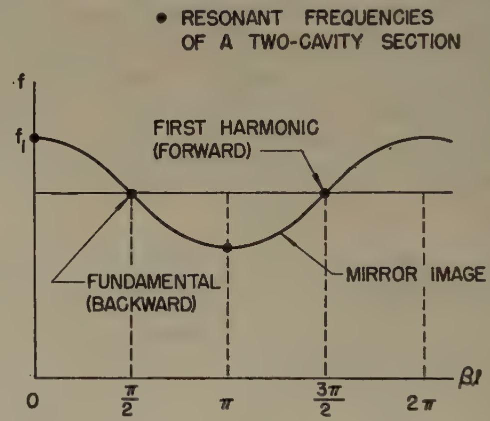

Fig. 3—Brillouin diagram (f vs. βl) for structure shown in Fig. 1.

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[4] See for example, J. C. Slater, "The design of linear accelerators," *Rev. Mod. Phys.*, vol. 20, p. 495; July, 1948.

short circuit, can be obtained directly from the Fourier analysis of Fig. 2 for the desired space harmonic m.

Perturbation Experiment

We now relate the total field E_t at the plunger to the energy stored in the multiple cavity, by means of a perturbation experiment. This part is closely related to the measurement of R/Q of a cavity.[5] The experiment consists of moving a small cylindrical plunger at the short circuit, causing both positive and negative perturbations of volume $\Delta\tau$ and recording the corresponding changes in resonant wavelength $\Delta\lambda$. If we consider the slope of this curve at zero perturbation, $(\Delta\lambda/\Delta\tau)_{\Delta\tau\to 0}$, and assume that the perturbation is sufficiently small so that H is essentially zero and E is constant $(E=E_t)$ over its surface, we can write (4) in the form

$$\left(\frac{E_t^2}{\text{Energy stored in cavity}}\right)=\frac{4}{\lambda\epsilon}\left(\frac{\Delta\lambda}{\Delta\tau}\right)_{\Delta\tau\to 0}$$

which can be also written in the form

$$\left(\frac{E_t^2 l^2}{c w_c}\right)=480\pi\,\frac{Ll^2}{\lambda}\left(\frac{\Delta\lambda}{\Delta\tau}\right)_{\Delta\tau\to 0}\quad\text{ohms}\qquad(7)$$

where:

E_t = field at the perturbation,

l = periodic length,

L = total length of the cavity,

w_c = energy stored per unit length *in the cavity*,

c = velocity of light,

$w_c L$ = energy stored in cavity,

$1/c\epsilon = 377$ ohms.

The value of the constant in (7) is correct only for those shapes of the perturbing plunger which do not alter the fields at large distances away. A cylindrical plunger where $\Delta\tau$ is made to approach zero by shrinking its height satisfies this condition. Since the energy stored per unit length in a traveling wave is given by $w=2w_c$, and the measurement of $(\Delta\lambda/\Delta\tau)_{\Delta\tau\to 0}$ was done inside a cavity (i.e., with standing waves), one can write

$$\left(\frac{E_t^2 l^2}{cw}\right)=240\pi\,\frac{Ll^2}{\lambda}\left(\frac{\Delta\lambda}{\Delta\tau}\right)_{\Delta\tau\to 0}\quad\text{ohms}\qquad(8)$$

which relates the field at the perturbation to the energy stored in a traveling wave.

Circuit Impedance

It now remains to tie all the results together. From the definition of circuit impedance [(2)], and after rearrangement,

$$K=\frac{E_m^2}{2\beta_m^2 w v_g}=\frac{1}{2}\left[\left(\frac{c}{v_g}\right)\left(\frac{1}{\beta_m l}\right)^2\right]_{(a)}\left[\frac{E_m}{E_t}\right]^2_{(b)}\left[\frac{E_t^2 l^2}{cw}\right]_{(c)}.\qquad(9)$$

Bracket (a) is obtained from the Brillouin diagram, and bracket (b) is determined from Fig. 2 by Fourier analysis of the axial E field. Bracket (c) is obtained from (8), by means of a perturbation experiment.

Thus all the factors are known and the circuit impedance can be calculated from cold measurements on a resonant section of the structure. The variation of impedance with frequency can also be obtained by repeating the experiment at other resonant frequencies of the multiple cavity section. The gain parameter C for a given beam current and voltage is then obtained by direct substitution in (1).

EXPERIMENTS

Table I summarizes the results of measurements on a disk-loaded type structure suitable for a space harmonic amplifier.

TABLE I

COMPARISON OF EXPERIMENTAL AND PREDICTED RESULTS

Column	1	2	3	4
βl	$\frac{E_m}{E_t}$	$\frac{E_t^2 l^2}{w_c}$	$\frac{E_m^2 l^2}{w_c}$	$\frac{E_m^2 l^2}{w_c}$
0	0.71	864	435	529
$\pi/2$	0.67	660	296	295
π	0.47	705	155	136
$3\pi/2$	0.30	660	60	90
2π	0.10	864	9	20
$5\pi/2$	0.05	660	1.6	2.3

Columns 1 and 2 are obtained as described earlier.

Column 4 gives a theoretical figure obtained from Pierce's "Pillbox" resonator calculation and transit time considerations[6] for a structure closely resembling the one used.

Agreement between columns 3 and 4 gives an indication of what can be expected, keeping in mind that the theroetical calculations of column 4 apply to a somewhat different structure.

CONCLUSION

A method has been outlined whereby the circuit impedance of periodically loaded structures, which can be made resonant by placing short circuits at symmetry planes, can be evaluated from cold tests on a sample cavity. The results agree with measured values of gain on a disk-loaded type TW tube whose behavior was predicted by this method, as well as with theoretical calculations on a disk-loaded structure in which the field distribution is reasonably well known. The method can also be extended to periodic structures in which there are no symmetry planes where short circuits can be placed. In this case one essentially ignores the fields in the vicinity of the end sections, and Fourier analyzes only the central portion of the fields. A sufficient number of sections must be used, so that the frequency of the mode is not perturbed by the shorts.

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[5] W. W. Hansen and R. F. Post, "On the measurement of cavity impedance," *Jour. Appl. Phys.*, vol. 19, pp. 1059–1061; November, 1948.

[6] J. R. Pierce, "Traveling-Wave Tubes," D. Van Nostrand Company, Inc., New York, N. Y., pp. 73, 93, 234: 1950.

A Simple Microwave Correlator*

R. H. WILCOX†, ASSOCIATE, IRE

Summary—**A simple correlator employing the ring modulator circuit as a rapid-response multiplier has been developed which accepts signals on coaxial cables and which is usable over a wide range of microwave frequencies. The correlator output includes a bias term, but the value of this term can be predicted mathematically so that the correlator may be used for quantitative measurements.**

INTRODUCTION

THE RECENT interest shown in electronic computing and analysis has considerably increased the demand for devices which will manipulate voltages and currents according to particular mathematical processes. In particular, the exploitation of the auto- and cross-correlation functions[1] (which require delay, multiplication, and averaging, in that order) as powerful means of studying signal characteristics has called for manipulating devices with unusually broad capabilities. The averaging is easily performed with low-pass filters, the delay can be achieved with real or artificial transmission lines or record-playback techniques, but the multiplication of two signals poses difficult problems whenever the frequencies involved are not low.

Most multiplying computers employ electromechanical servomechanisms, but these are inherently limited to very low-frequency response. All-electronic multipliers involving multigrid vacuum tubes have been developed, but they are complex, subject to drift, and limited to frequencies at which the drift time of the electron stream through the tube elements is of no consequence, i.e., to a few hundred megacycles. Obviously, for a system to perform successful multiplication at microwave frequencies it should be essentially insensitive to frequency variations (no tuned elements, no critical drift times, and wide-band impedance matches); it should be capable of receiving signals from coaxial transmission lines; and it should be as simple as possible, in order to reduce drift and critical adjustments. A microwave correlation system has been developed[2] which employs frequency translation and compression to present the original microwave information at lower frequencies, where it is easily treated with conventional circuitry. However, if the input signals are of sufficient strength, they may be correlated at the original microwave frequencies by means of the simpler passive device to be described.

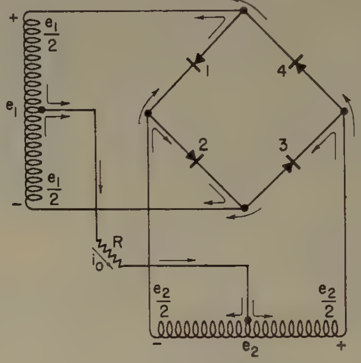

Fig. 1—Basic multiplier circuit.

THE BASIC MULTIPLIER CIRCUIT

The well-known ring modulator circuit, when used as shown in Fig. 1[3], forms a simple multiplying bridge[4] in which the current i_0 flowing through the load resistor R is proportional to the product of the two applied voltages e_1 and e_2. This hypothesis can be explained by the following method, in which it is assumed that:

a. The current through a crystal diode is a fixed exponential function of the voltage across it; i.e., $i = A(\epsilon^{\gamma V} - 1)$, in which A and γ are constant regardless of voltage polarity (this is true for small values of current and voltage).
b. The source impedances are negligible with respect to the crystal and load impedances (this can be made so).
c. The four diodes in the bridge have identical current-voltage characteristics (this condition can be approached by careful selection).

The current i_0 through the load resistor R must be the sum of individual currents following each of four possible paths, one through each diode. The voltage V across each crystal must be equal to the sum of the voltages around the current path associated with that crystal, according to Kirchoff's voltage law. Thus

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* Decimal classification: 621.375.2. Original manuscript received by the IRE, March 10, 1954; revised manuscript received, June 21, 1954. The opinions or assertions contained in this manuscript are those of the writer and are not to be construed as official or reflecting the views of the Navy Department or the Naval Service as a whole.

† Naval Research Laboratory, Washington, D. C.

[1] H. M. James, N. B. Nichols, and R. S. Phillips, "Theory of Servomechanisms," Radiation Laboratory Series, McGraw-Hill Book Company, Inc., New York, N. Y., vol. 25, pp. 273–277; 1947. J. L. Lawson and G. E. Uhlenbeck, "Threshold Signals," Radiation Laboratory Series, McGraw-Hill Book Company, Inc., New York, N. Y., vol. 24, pp. 33–39, 1950.

[2] R. M. Page, A Brodzinsky, and R. R. Zirm, "A microwave correlator," PROC. I.R.E., vol. 41, pp. 128–131; January, 1953.

[3] It should be noted that the polarities and directions indicated in Fig. 1 are for the sole purpose of facilitating consistant analysis and do not necessarily bear any relation to actual voltage polarities or current directions.

[4] R. H. Wilcox, "Crystal Diode Ring Multipliers," NLR Report 4385, June, 1954.

$$i_0 = i_1 + i_2 + i_3 + i_4, \tag{1}$$

but

$$i_1 = +A(\epsilon^{\gamma V_1} - 1) = +A \sum_{n=1}^{\infty} \frac{(\gamma V_1)^n}{n!}$$

$$= +A \sum_{n=1}^{\infty} \frac{\gamma^n \left(+ \frac{e_1}{2} + i_0 R + \frac{e_2}{2} \right)^n}{n!}$$

$$i_2 = -A(\epsilon^{\gamma V_2} - 1) = -A \sum_{n=1}^{\infty} \frac{(\gamma V_2)^n}{n!}$$

$$= -A \sum_{n=1}^{\infty} \frac{\gamma^n \left(- \frac{e_1}{2} + i_0 R + \frac{e_2}{2} \right)^n}{n!}$$

$$i_3 = +A(\epsilon^{\gamma V_3} - 1) = +A \sum_{n=1}^{\infty} \frac{(\gamma V_3)^n}{n!}$$

$$= +A \sum_{n=1}^{\infty} \frac{\gamma^n \left(- \frac{e_1}{2} + i_0 R - \frac{e_2}{2} \right)^n}{n!}$$

$$i_4 = -A(\epsilon^{\gamma V_4} - 1) = -A \sum_{n=1}^{\infty} \frac{(\gamma V_4)^n}{n!}$$

$$= -A \sum_{n=1}^{\infty} \frac{\gamma^n \left(+ \frac{e_1}{2} + i_0 R - \frac{e_2}{2} \right)^n}{n!}. \tag{2}$$

Since the value of γ for microwave crystal diodes is about 15, for $V \leqq 0.1$, all terms containing $n > 4$ will be at least an order of magnitude smaller than the $n = 1$ term and can be neglected. Of the remaining terms in the four expansions, some cancel, some are sufficiently small to be neglected, and the rest sum to the expression:

$$i_0 = \left[A\gamma^2 + \frac{A\gamma^4}{24} (e_1^2 + e_2^2) \right] e_1 e_2. \tag{3}$$

ADAPTATION TO MICROWAVE OPERATION

In adapting the multiplying circuit shown in Fig. 1 to microwave operation, provision must be made to receive signals from coaxial transmission lines. However, if the two inputs e_1 and e_2 are applied to the bridge on coaxial cables, then one of the crystals will be shorted by the common ground of the cable outer conductors. If the input terminals marked (−) in Fig. 1 are connected to a common ground, then diode number 2 will be shorted, as shown in Fig. 2. Under this condition, the load current i_0 will include a current

$$i_5 = \frac{1}{R} \left(- \frac{e_1}{2} + \frac{e_2}{2} \right) = \frac{e_2 - e_1}{2R}, \tag{4}$$

which will replace i_2 in (1). When the remaining three exponential currents are summed with i_5 the unbalanced condition results in the presence of several significant terms that do not appear in (3):

$$\begin{aligned} i_0 = {} & A\gamma i_0 R + \frac{A\gamma^2}{8} (e_1^2 + e_2^2) + \frac{3A\gamma^2}{4} (e_1 e_2) \\ & + \frac{A\gamma^4}{32} (e_1^2 + e_2^2)(e_1 e_2) \\ & + \frac{A\gamma}{2} (e_2 - e_1) + \frac{A\gamma^2 i_0 R}{2} (e_2 - e_1) \\ & - \frac{A\gamma^3 e_1 e_2}{16} (e_2 - e_1) + \frac{e_2 - e_1}{R}. \end{aligned} \tag{5}$$

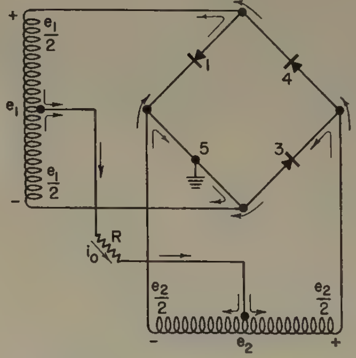

Fig. 2—Unbalanced multiplier circuit.

When this unbalanced bridge is followed by a low-pass filter, as is required for correlation, all the terms containing odd powers of e_1 or e_2 will average to zero, leaving only terms containing the product $(e_1 e_2)$ or the quantity $(e_1^2 + e_2^2)$:

$$i_0 = \frac{A\gamma^2 [24 + \gamma^2 (e_1^2 + e_2^2)]}{32(1 - A\gamma R)} (e_1 e_2) + \frac{A\gamma^2 (e_1^2 + e_2^2)}{8(1 - A\gamma R)}. \tag{6}$$

The first, or product, term is a factor which is proportional to the average of the instantaneous product of e_1 and e_2; the second, or bias, term is a parameter which is proportional to the average values of e_1^2 and e_2^2. This bias parameter has the effect of displacing the output function from its normal axis; if the average values of e_1^2 and e_2^2 are the same (as when determining an autocorrelation function) and do not exceed 0.1 volt then the average value of the bias term is approximately one-third the maximum average value of the product term. It will be seen later that the displacement obtained experimentally under these conditions agrees quite well with this predicted value.

For practical operation of this unbalanced bridge the experiment model was constructed. (Fig. 3, page 1514.) Resistances R_T served to terminate the input transmission lines (RG-8/U cable) in their characteristic impedance (52 ohms). Originally, four quarter-watt carbon resistors were located symmetrically between the center and outer conductors, but it has since been found that a painted carbon film across the face of the dielectric in a receptacle connector (UG-58/U) results in an even lower VSWR and a wider bandwidth. The protruding center conductors were then connected through the blocking capacitors C_B (100 $\mu\mu f$) to the unbalanced bridge, in which germanium diodes (1N72) were employed. The center-tapping resistors R_D (4700 Ω) were incorporated to locate the electrical mid-points of the input signals. In the original model, coaxial connectors were included at points A and A', so that a fourth diode could be added and comparisons made between balanced and unbalanced operation. Since four perfectly matched diodes are difficult to obtain, some displacement of the output axis was found even with the balanced bridge. This experimentally determined "balanced axis" was used as the zero axis for the measurement of displacement due to unbalancing the bridge.

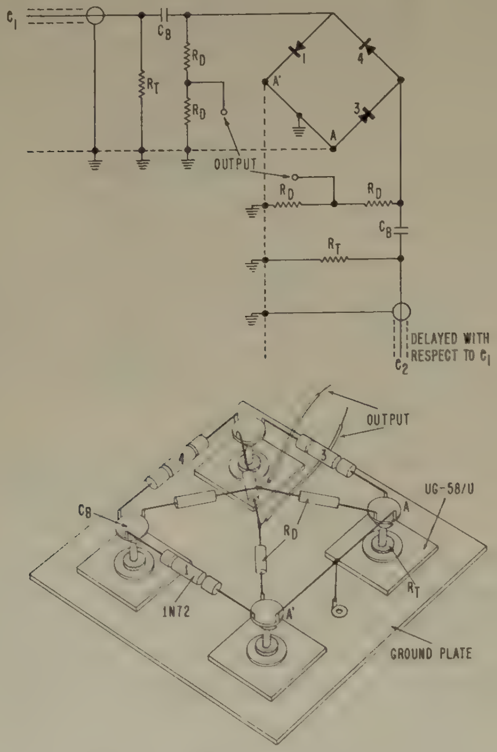

Fig. 3—Microwave correlator.

Experimental Results

The unbalanced bridge described above was used to measure the autocorrelation function of the 1.1-kMc sine wave. This signal was obtained from a signal generator, split with a coaxial "tee," and fed through two coaxial cables to the inputs of the unbalanced bridge. Delay was obtained by varying the length of one cable with respect to the other. The bridge output was averaged and measured with an optical galvanometer. This experimentally obtained function is compared with the theoretical autocorrelation function of a 1.1-kMc sine wave in Fig. 4. By displacing the axis of the theoretical function $33\frac{1}{3}$ per cent of its maximum amplitude (the predicted amount), good agreement of the two curves is obtained.

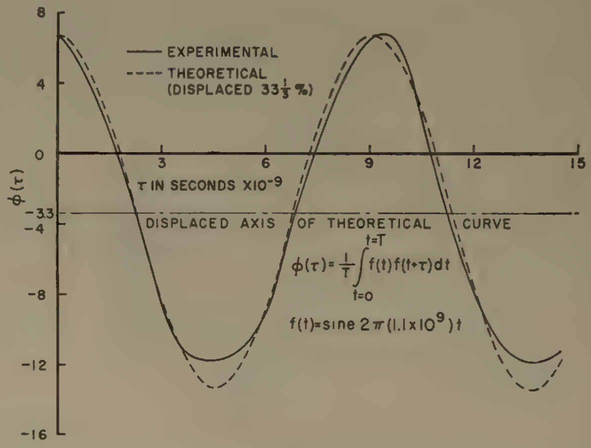

Fig. 4—Comparison of experimental and theoretical determinations of the autocorrelation function of a 1.1-kMc sine wave.

The signal generator was replaced with a random noise source whose output was roughly bandlimited between 1.0 and 1.2 kMc by filters. The autocorrelation function was measured as before and is shown in Fig. 5. It was assumed that the "balanced axis" remained the same as in the preceding experiment, and the predicted axis displacement of $33\frac{1}{3}$ per cent of the maximum amplitude was made from this predetermined level. The actual axis (the value at which the function went to zero) was displaced by 38 per cent from the balanced axis; the difference of 5 per cent could easily result from a lack of perfect crystal matching.

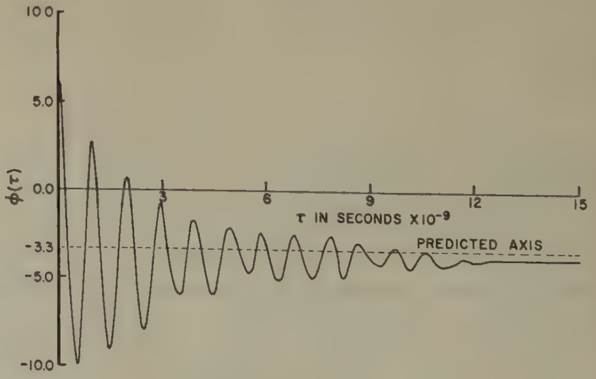

Fig. 5—Experimentally determined autocorrelation function of random noise bandlimited between 1.0 and 1.2 kMc.

Conclusion

The variation in current-voltage characteristics of available crystal diodes was found to be the greatest deterent to quantitative prediction of axis displacement.

This disadvantage is common to both the balanced and unbalanced bridges, but patient selection from a sufficiently large supply of diodes should reduce it to a minimum.

The dynamic power range of this correlator is relatively small. If the input signal strength exceeds about ±0.1 volt, the values of A and γ in the crystal characteristic change and distortion is introduced. Conversely, too small a signal is lost in the noise generated in the diodes. Also, since this correlator is a passive device, the output is of small magnitude (a few microamperes) and must be measured with a sensitive galvanometer or amplified with a differential dc amplifier.

In the analysis of the unbalanced bridge operation it was shown that the output must be averaged over an interval which is long with respect to the input voltage variations, if the nonproduct terms in the output are to appear as a bias. Thus the unbalanced bridge is a rapid-response multiplier, not an instantaneous multiplier. This limitation is no disadvantage, when bridge is included with prior delay and post averaging in a correlator.

A correlator including this unbalanced bridge multiplier is usable over a very wide frequency range; the limits are those of the input cables themselves and their terminations. It is simple in construction and operation, and the constituent parts are not given to drift over long periods of time.

ACKNOWLEDGMENT

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The Calibration of Amplitude Modulation Meters with a Heterodyne Signal[*]

J. ROSS MACDONALD[†], SENIOR MEMBER, IRE

***Summary*—Errors which may occur in the calibration of peak or average reading amplitude modulation meters with a heterodyne signal are investigated in detail. Relations between the heterodyne-amplitude ratio *M* and the true modulation factor *m* of a sinusoidally modulated wave are derived. Using these results, heterodyne calibration of modulation analyzers to read *m* exactly can readily be carried out.**

INTRODUCTION

AMPLITUDE-MODULATION ANALYZERS for intermodulation measurements at audio frequency[1,2] or for the determination of modulation depth in radio transmission[3] are commonly calibrated with a heterodyne signal produced by summing two signals of different amplitudes and frequencies whose frequency difference is small compared to their frequencies. The reason for the use of such a calibration signal in place of a truly sinusoidally modulated signal is that the latter is difficult to produce with known modulation factor *ab initio*, whereas the former is easily produced with an accurately known ratio between the amplitudes of the beating signals. Since it is apparently not generally recognized that appreciable errors can arise when this calibration procedure is employed, this paper discusses how such errors occur and how they can be eliminated.

PRINCIPLES OF MODULATION MEASUREMENT

Most modulation analyzers read directly in terms of modulation factor and to do so perform the following operations on an input signal. First, the signal passes through a high-pass filter which removes any low-frequency modulating component which may be present and leaves the higher-frequency modulated carrier intact. In order to allow a final indication in terms of modulation factor alone, the average amplitude of the modulated carrier is set to a given level determined during initial calibration of the instrument. After necessary amplification, the modulated carrier is then demodulated with either a half- or full-wave linear detector. The rectified output of the detector then passes through a low-pass filter which eliminates unmodulated signal components at the carrier frequency and higher. The signal at the output of this filter then consists of the low-frequency modulating component of the modulated wave together with a dc component which is next eliminated by a series capacitor.

The resulting output signal is the low-frequency modulation alone. By comparing its amplitude with that of the carrier, the modulation factor may be directly obtained. In intermodulation testing, the modulating signal is usually rectified with a full-wave linear rectifier and its rms amplitude read on a dc meter, which responds to the average value of the rectified signal. On the other hand, modulation analyzers for rf signals commonly provide means to measure the amount of positive and negative peak modulation separately. They therefore use a half-wave detector and peak reading dc output meter which can be switched to read either positive or negative peaks of the modulating signal.[3]

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[*] Decimal classification: R254.111. Original manuscript received by the IRE, April 16, 1954; revised manuscript received, May 20, 1954.

[†] Texas Instruments Inc., Dallas, Texas

[1] J. K. Hilliard, "Distortion tests by the intermodulation method," PROC. I.R.E., vol. 29, pp. 614–620; December, 1941.

[2] J. M. van Beuren, "Simplified intermodulation measurements," *Audio Engineering*, vol. 34, p. 24; November, 1950.

[3] F. E. Terman, "Radio Engineers' Handbook," First Edition, p. 987, McGraw-Hill, New York, N. Y.; 1943.

ANALYSIS OF HETERODYNE CALIBRATION

In this section, we shall show that errors can arise from calibration with a heterodyne signal whether an output meter measuring the peak or average of the rectified-modulating signal is employed. The errors are appreciably smaller in the latter case but arise in both cases from the difference in waveshape between a true sinusoidally-modulated wave and a heterodyne wave.

The equation for a sinusoidally modulated wave may be written

$$\begin{aligned} e_0(t) &= A[1 + m \cos \omega_m t] \cos \omega_c t \\ &= A\left[\cos \omega_c t + \frac{m}{2} \cos (\omega_c + \omega_m)t \right. \\ &\quad \left. + \frac{m}{2} \cos (\omega_c - \omega_m)t\right], \end{aligned} \tag{1}$$

where A is a constant and m is the modulation factor, which ranges between zero and unity. The modulation analyzer separates out the modulating signal

$$e_m(t) = Am \cos \omega_m t, \tag{2}$$

and the output reading is proportional to its peak amplitude Am (the same for either positive or negative peaks) or its average amplitude, $2Am/\pi$ for full-wave rectification, Am/π for half-wave rectification. Assuming that the modulation factor m of the above signal is known, we shall first discuss how the modulation analyzer can in theory be calibrated to read m directly, then compare the procedure necessary when a heterodyne signal instead of a sinusoidally modulated signal is used for such calibration.

First, the signal of (1) is applied to the analyzer, whose gain is then adjusted so that the modulation meter reading corresponds to the known m. With the same gain, a reading R proportional to the average value of the modulated carrier (1) after rectification is then obtained either with a separate meter or by bypassing the first detector and reading the result on the modulation meter. Since this average value is simply A, independent of m, the analyzer will now yield the correct value of m for any other modulated signal like (1) but with different A and m, so long as the gain of the analyzer is first adjusted to give the same reading R for the average value of the rectified-modulated carrier. We see that this calibration procedure effectively removes the factor $2A/\pi$ or A/π occurring in the average value of the rectified-modulating signal by normalizing the average amplitude of the modulated carrier to a given value. A similar procedure may be used for a peak-reading output meter.

A pure, sinusoidally modulated signal of known modulation factor is seldom available for calibration purposes. On the other hand, however, a heterodyne signal $e_h(t)$ may be readily formed by adding the two signals $e_1(t) = A \cos \omega_c t$ and $e_2(t) = AM \cos (\omega_c + \omega_m)t$ to yield

$$\begin{aligned} e_h(t) &= A\left[\cos \omega_c t + \frac{M}{2} \cos (\omega_c + \omega_m)t \right. \\ &\quad \left. + \frac{M}{2} \cos (\omega_c + \omega_m)t\right] \\ &= A\left[(1 - M) \cos \omega_c t \right. \\ &\quad \left. + 2M \cos \left(\frac{\omega_m t}{2}\right) \cos \left(\frac{2\omega_c + \omega_m}{2}\right) t\right], \end{aligned} \tag{3}$$

where M is the ratio of the smaller signal e_2 to the larger, e_1.[4] M may range between zero and unity, and the incorrect assumption is often made that it is equal to the modulation factor m for a true sinusoidally-modulated wave. Comparison of the second form of (1) and the first form of (3) shows, however, that although the signals are very similar, the modulated wave has upper and lower sidebands, whereas the heterodyne wave has only an upper (or lower) sideband. We shall now derive relations between M and m for positive or negative peak or for average-reading modulation meters which will allow correct calibration to be achieved with a heterodyne-calibrating signal for any value of M.

In order to obtain relations between M and m, it is necessary to investigate in detail what happens to the heterodyne signal as it passes through the various stages in the modulation analyzer. Fig. 1, on the next page, gives a comparison of the forms of a sinusoidally modulated wave with $m = 1$ in (a) and a heterodyne wave with $M = 1$ in (b). The difference between these waveshapes is very small for m and $M \ll 1$, and is a maximum when these quantities equal unity. After passing through a linear half- or full-wave detector and a low-pass filter, the waveform in (c) is obtained. The filter must, of course, have a sufficiently wide pass band that the sharp minima in (c) be passed with negligible distortion, yielding an accurate reproduction of the positive envelope of the heterodyne wave. Its equation for arbitrary M, which is readily obtained by a vector addition of $e_1(t)$ and $e_2(t)$, is

$$e_e(t) = [A^2 + M^2A^2 + 2MA^2 \cos \omega_m t]^{1/2}. \tag{4}$$

If we let $\omega_m t = 2\theta$ and $k^2 = 4M/(1+M)^2$, (4) may be rewritten as

$$e_e(\theta) = A(1 + M)[1 - k^2 \sin^2 \theta]^{1/2}. \tag{5}$$

The average value of this envelope is

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[4] M.I.T. Electrical Engineering Staff, "Applied Electronics," p. 699, John Wiley and Sons, Inc., New York, N. Y.; 1943.

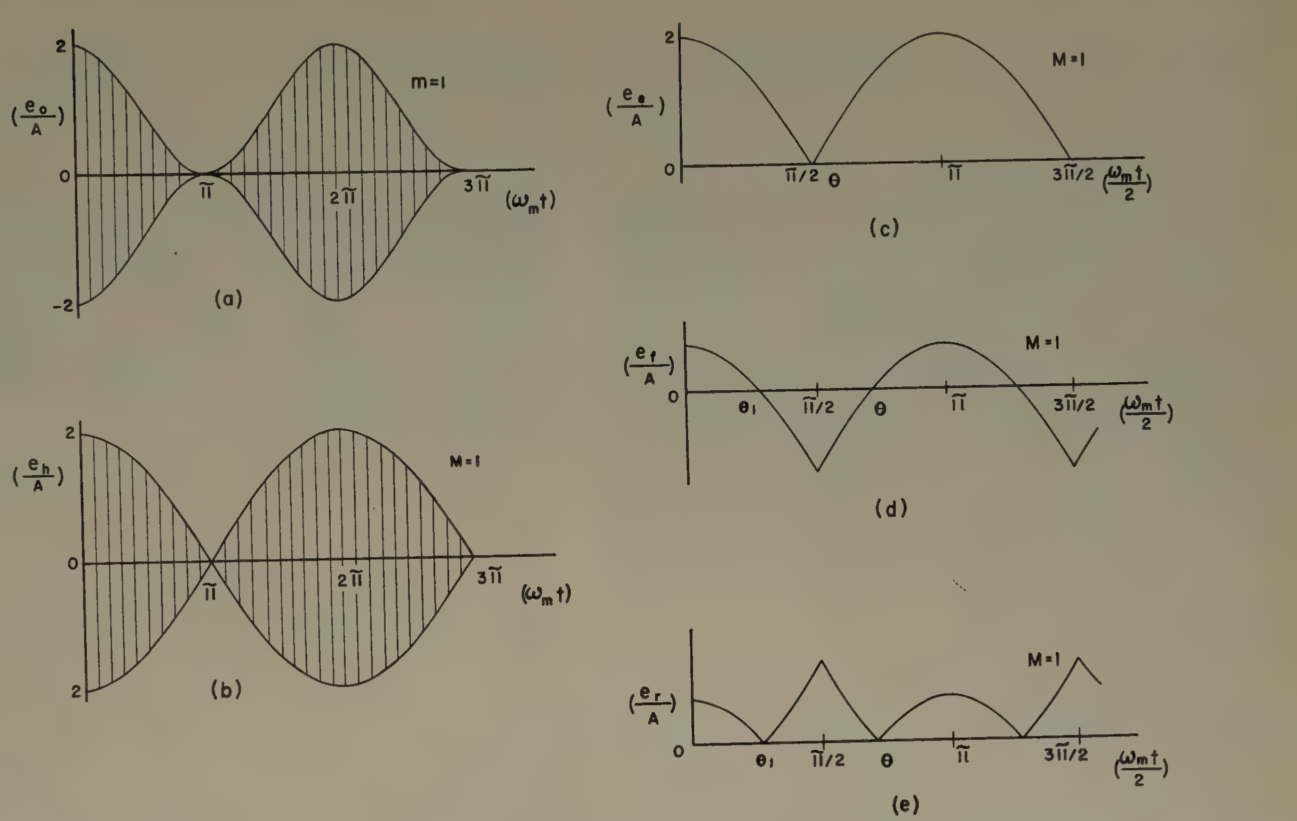

Fig. 1—Waveshapes in modulation analyzer. (a) Sinusoidally modulated input signal. (b) Heterodyne input signal. (c) Envelope of rectified heterodyne signal. (d) Envelope of rectified heterodyne signal with dc component removed. (e) Full-wave rectified waveshape of signal in (d).

$$\bar{e}_e = \frac{2A(1+M)}{\pi}\int_0^{\pi/2}[1-k^2\sin^2\theta]^{1/2}d\theta$$
$$= \frac{2A(1+M)}{\pi}E, \qquad (6)$$

where E is the complete elliptic integral of the second kind. Note that as M approaches zero, $2E/\pi$ approaches unity and $\bar{e}_e$ approaches A, the value obtained for a sinusoidally modulated wave. For $M>0$, however, $2E/\pi$ is a function of M and is less than unity.

When the waveform (5) is passed through a condenser, the dc component is eliminated and the equation for the resulting signal $e_f(\theta)$ is

$$e_f(\theta) = A(1+M)[\sqrt{1-k^2\sin^2\theta} - 2E/\pi]. \qquad (7)$$

The waveshape of $e_f(\theta)$ for $M=1$ is shown in Fig. 1(d). It is now necessary to determine the value of θ for which $e_f(\theta)=0$; this value, θ_1, is readily found to be

$$\theta_1 = \sin^{-1}\left\{\frac{1}{k}[1-(2E/\pi)^2]^{1/2}\right\}. \qquad (8)$$

θ_1 approaches $\pi/4$ as M tends to zero and increases slightly above this value for nonzero M. The positive and negative peaks of the waveform of (7) occur at $\theta=0$ and at $\theta=\pi/2$ and are

$$e_f(0) = A(1+M)(1-2E/\pi) \qquad (9)$$

$$e_f(\pi/2) = A(1+M)(\sqrt{1-k^2} - 2E/\pi). \qquad (10)$$

If the signal of (7) is full-wave rectified, the waveshape between $\theta=0$ and $\theta=\pi/2$ is given by

$$\begin{aligned} e_r(\theta) &= e_f(\theta) \quad 0 \leqq \theta \leqq \theta_1 \\ &= -e_f(\theta) \quad \theta_1 \leqq \theta \leqq \pi/2, \end{aligned} \qquad (11)$$

and is shown in Fig. 1(e) for $M=1$. A dc meter will respond to the average value of this rather peculiar wave. This average value is

$$\bar{e}_r = \frac{2A(1+M)}{\pi}\int_0^{\pi/2} e_r(\theta)d\theta$$
$$= \frac{4A(1+M)}{\pi}\int_0^{\theta_1} e_f(\theta)d\theta. \qquad (12)$$

The last equation follows because the area under the curve from $\theta=0$ to θ_1 is necessarily equal to that from $\theta=\theta_1$ to $\pi/2$. The result of the above integration is

$$\bar{e}_r = [4A(1+M)/\pi][E(\theta_1, k) - 2\theta_1 E/\pi], \qquad (13)$$

where $E(\theta_1, k)$ is the incomplete elliptic integral of the second kind.

We are now finally in a position to obtain the relation between the true modulation factor m and the heterodyne signal ratio M. First, we shall consider an analyzer which measures the average value of $e_r(\theta)$. Part of the calibration procedure, as discussed above, involves adjusting the gain of the analyzer so that the rectified-modulated carrier gives a certain set reading on a dc meter. For a sinusoidally modulated wave, this reading is independent of m, but for a heterodyne wave it depends on M, as shown by (6). This dependence can introduce error of its own into the calibration with a heterodyne signal unless it is properly taken into account. Because of this potential error and because of differences between the waveshape of a sinusoidally modulated wave and of a heterodyne wave, m is not equal to M. We define $\bar{m}$ in the present average-reading case as the true modulation factor for a sinusoidally modulated wave corresponding to a given M for a heterodyne wave. The calibration procedure using a heterodyne signal therefore consists of adjusting the analyzer gain so that the heterodyne signal gives a modulation reading of $\bar{m}$ (not M) and then reading a quantity R proportional to the average value of the rectified-modulated carrier $\bar{e}_e$. After such calibration, the analyzer will measure the true modulation factor m for a sinusoidally modulated signal when the carrier amplitude is initially set to R.

For the above type of analyzer which measures a quantity proportional to $\bar{e}_r$, the actual modulation-meter reading, after setting the rectified modulated-carrier-level to the proper value will be $\bar{m}$, by definition, where $\bar{m}$ is given by $\bar{e}_r/\bar{e}_e$. We therefore obtain

$$\bar{m} = (2E/\pi)^{-1}[2E(\theta_1, k) - 4\theta_1 E/\pi]. \qquad (14)$$

The value of M corresponding to $\bar{m}$ in (14) is, of course, implicit in E, $E(\theta_1, k)$, k, and θ_1.

For an analyzer which measures positive and negative peaks of the modulating voltage $e_f(\theta)$, the actual modulation-meter readings, again after setting the proper modulated-carrier level, are m^+ for positive peaks, m^- for negative, where these quantities are given by $e_f(0)/\bar{e}_e$ and $-e_f(\pi/2)/\bar{e}_e$, respectively. The results are

$$m^+ = (2E/\pi)^{-1} - 1 \qquad \text{(positive peak reading)} \quad (15)$$

$$m^- = 1 - \sqrt{1 - k^2}\,(2E/\pi)^{-1} \quad \text{(negative peak reading)}. \quad (16)$$

Eqs. (14), (15), and (16) are the most important results of this analysis and allow the different types of modulation meters to be calibrated to read m correctly with a heterodyne wave. In Table I we tabulate the dependence of $\bar{m}$, m^+, and m^- on M for $M=0$ to 1. The percentage differences of these quantities from M are plotted versus M in Fig. 2. These calculations were carried out using the tables of elliptic functions given by Jahnke and Emde.[5] Fig. 2 indicates that less than one per cent error in calibration will occur when M is taken equal to m in the range $M<0.04$ for m^+ and m^-, and $M<0.17$ for $\bar{m}$. In addition, a negative peak reading meter can also be calibrated with less than one per cent error with M taken equal to m^- when $0.96<M\leqq 1$. On the other hand, if meters reading m^+ or $\bar{m}$ are calibrated near $M=1$, the error in a modulation reading near $M=1$ will be of the order of 30 or 40 per cent if M is erroneously assumed to equal m.

TABLE I

Dependence of $\bar{m}$, m^+, and m^- on Heterodyne Signal Ratio M

M	$\bar{m}$	m^+	m^-
0	0	0	0
0.007654	0.00765	0.007628	0.007680
0.03110	0.0311	0.03084	0.03133
0.07180	0.07160	0.07039	0.07302
0.1325	0.1316	0.1276	0.1362
0.2174	0.2136	0.2032	0.2266
0.3333	0.3183	0.2970	0.3515
0.4903	0.4467	0.4045	0.5196
0.7041	0.5796	0.5102	0.7378
1	0.6613	0.5708	1

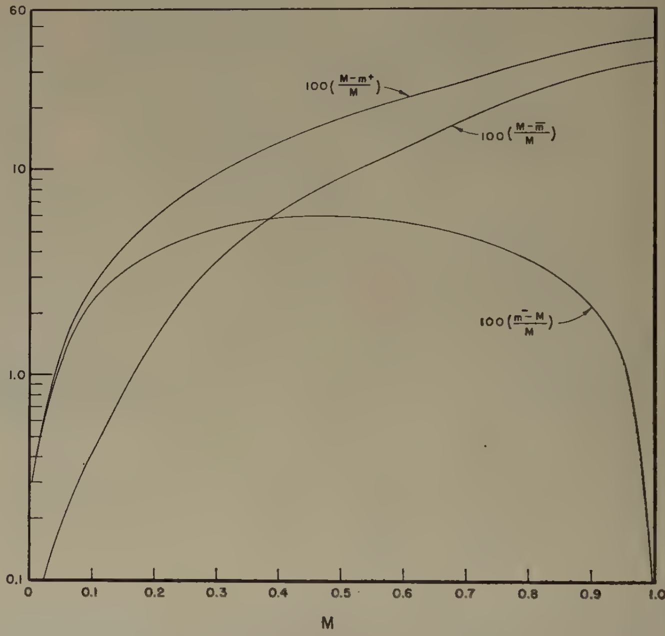

Fig. 2—Percentage difference between true modulation factors $\bar{m}$, m^+, and m^- and heterodyne amplitude ratio M as functions of M.

CONCLUSIONS

The calibration of amplitude-modulation analyzers with a heterodyne-calibrating signal has been analyzed in detail and potential errors implicit in such calibration have been investigated. It has been found that the error may be as large as 40 per cent if the heterodyne-amplitude ratio M is taken equal to the true modulation factor m of a sinusoidally modulated wave when M is near unity.

Essentially errorless calibration with a heterodyne signal can be carried out in either of two ways. First, if M and m are assumed equal but a value of M much less than unity is used, negligible error will result. A second method, which is exact for any value of M, requires only that the value of m obtained either from Table I or calculated from the formulae derived in this paper be used in the calibration procedure, instead of M.

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[5] E. Jahnke and F. Emde, "Tables of Functions," Dover, N. Y.; 1943.

A New Transducer Diagram*

R. N. BRACEWELL†

***Summary*—A simple diagram consisting of two graduated straight lines and a point, completely represents a loss-free transducer. By a simple projective construction the input reactance corresponding to any terminating reactance may be quickly obtained, and, by an extension of the construction, complex impedances may also be transformed. The method allows a visual as well as numerical grasp to be obtained of the transforming action of the transducer. The construction depends on the principle that in any bilinear transformation of a function of a complex variable there are straight lines which transform into straight lines, and that the relation between corresponding points on these straight lines is projective.**

INTRODUCTION

THE WELL KNOWN circle diagrams of Kennelly and Smith have proved extremely useful in handling problems involving the transformation of impedances by lengths of transmission line. To a great extent their value stems from the graphical nature of the procedure. It enables a quick and clear grasp of the transformation to be obtained, which is not afforded by the numerical methods using complex numbers. The present communication, which deals with a more general problem than that of the uniform transmission line, presents a graphical procedure for transforming impedances through an arbitrary loss-free transducer.

For any such transducer there exists a particularly simple diagram (Fig. 1) based on ideas of projective geometry, which, when drawn, allows the input impedance corresponding to any terminating impedance to be quickly deduced.

To the knowledge of the writer, the only projective diagram previously published for performing this transformation is that of Weissfloch.[1] The latter is considerably more cumbersome to use and more susceptible to graphical errors than the one described below. The relation between the two diagrams is demonstrated later. The present diagram was originally described in a restricted wartime report.[2]

ESTABLISHING THE DIAGRAM

In order to specify a transducer completely, three independent measurements must be made on it. Let the transducer be terminated with three different reactances a, b, c, in turn and let the input reactances be measured. Let these be A, B, C, respectively. In practice the terminating reactances would often include 0 and ∞, i.e., measurements would be made with the transducers short-circuited and open-circuited.

Now let the quantities a, b, c, A, B, C, be marked off on two linear scales whose scale factors and origins may be chosen at will to suit the values of the quantities. Lay one scale on the other, at any convenient angle, so that A and a coincide (Fig. 1). Then mark the point P which lies on the intersection of Bb and Cc. The two graduated scales and the point P then comprise the desired diagram.

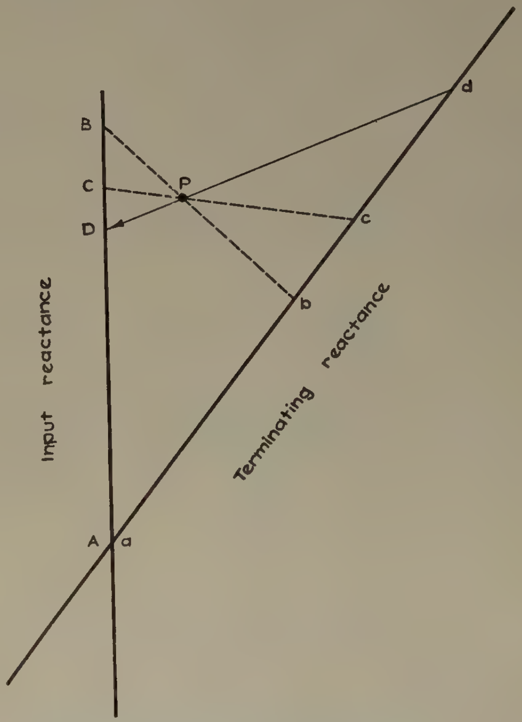

Fig. 1—The transducer diagram.

We shall now prove that if the transducer is terminated with any reactance d, then the input reactance is D, where D is the intersection of dP with AB. We shall assume without proof the following proposition from projective geometry.

If the points of two straight lines α and β are in (1, 1) correspondence, the general analytical relation between co-ordinates x_α and x_β on the two lines is bilinear, i.e., of the form

$$x_\beta = \frac{\phi + \chi x_\alpha}{\psi + \omega x_\alpha},$$

where ϕ, χ, ψ, ω are real constants.

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* Decimal classification: R145. Original manuscript received by the IRE, April 29, 1954.

† Radiophysics Lab., Commonwealth Scientific and Industrial Research Organization, Sydney, Australia.

[1] A. Weissfloch, "Kreisgeometrische Vierpoltheorie und ihre Bedeutung für Messtechnik und Schaltungstheorie des Dezimeter- und Zentimeterwellengebietes," *Hochfrequenz. und Elektroakustik*, vol. 61, pp. 100–123; April, 1943.

[2] R. N. Bracewell, "A graphical means of investigating loss-free transducers," Radiophysics Laboratory Report T.I. 226, 1945.

Now from the theory of linear transducers it is known that the input impedance Z_1 is a bilinear function of the terminating impedance Z_2 with three independent complex constants, i.e.,

$$Z_1 = \frac{c + dZ_2}{a + bZ_2},$$

where a, b, c, d, are the "linear parameters" as customarily defined. As a particular case of this, the input impedance to a loss-free transducer terminated in a reactance X_2 is itself reactive $(=jX_1)$, and the relation between the reactances X_1 and X_2 is bilinear with three independent real constants. But this is a projective relationship, and therefore if it is made to agree with a particular projective transformation in three independent instances, then it must do so in all. Hence the points d and D in Fig. 1 represent corresponding reactances.

Although the data specifying the transducer have been presented in the form of electrical measurements of a certain kind, the transformation diagram may also be set up from any other specification of the transducer such as the equivalent-T impedances, or short and open circuit impedances; and conversely, beginning from the diagram, any other specification may be deduced. It is sufficient to note the simple constructions which give the open and short-circuit impedances Z_{o1}, Z_{o2}, Z_{s1}, Z_{s2} (Fig. 2).

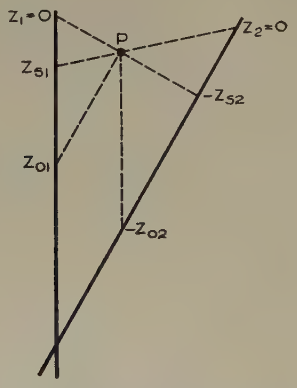

Fig. 2—The short and open circuit impedances are obtained by joining P to the zero points and points at infinity on the two reactance scales.

TRANSFORMATION OF COMPLEX IMPEDANCES

It remains to be shown how complex impedances may also be transformed by the diagram, and for this purpose we shall need the following two results from the theory of functions of a complex variable.

(1) The transformation $w=f(z)$, where w is an analytic function of the complex variable z, is conformal; i.e. angles are preserved.

(2) The conformal transformation represented by the bilinear relation

$$w = \frac{\alpha + \beta z}{\gamma + \delta z},$$

transforms circles into circles. The constants α, β, γ, δ, may be complex.

Let the terminating impedance $Z_2=R_2+jX_2$ be represented by a point on the $O_2R_2X_2$ plane (Fig. 3). It is required to find the point Z_1 on the $O_1R_1X_1$ plane representing the corresponding input impedance. The points of the two planes are in (1, 1) correspondence, but the correspondence, unlike that between the two reactance axes, is not projective. Since Z_1 is a bilinear function of the complex variable Z_2, the correspondence is the conformal one in which circles correspond to circles. These remarks lead to the following direct procedure for getting from Z_2 to Z_1.

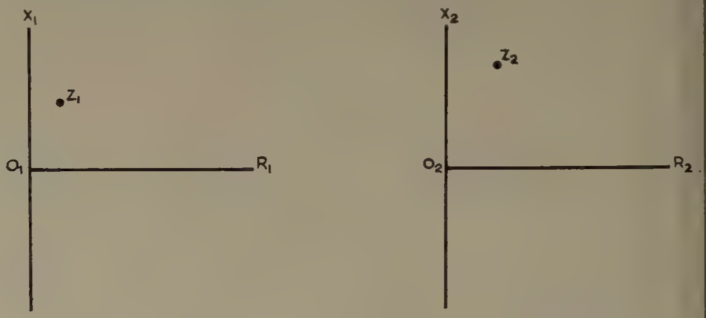

Fig. 3—The transformation of complex impedances.

Place the two reactance axes in the correct projective correspondence with respect to the point P (Fig. 4). Through Z_2, draw any two convenient semicircles, each with center on O_2X_2. (The construction has been simplified by making one of the semicircles of infinite radius.) Projectively transform the intersections of the circles with the O_2X_2 axis and construct two semicircles based on the points so obtained. The intersection of the two semicircles is the desired point Z_1.

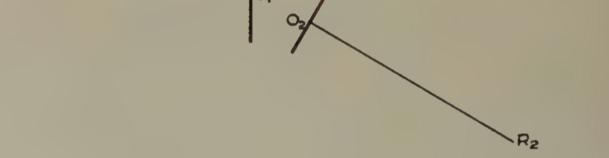

Fig. 4—Construction for the transformation of complex impedances

To prove this result we note that Z_1 must lie on the intersection of the transforms of the two circles through Z_2. These two circles must themselves transform into circles. Since the reactance axis O_2X_2 transforms into the axis O_1X_1 and since the first pair of circles cut O_2X_2 at right angles, it follows that the transformed circles must cut O_1X_1 at right angles. Four points on O_1X_1 are thus sufficient to construct the two transformed circles, and as they are on the reactance axis, they are within the scope of the projective transformation with respect to P.

The Dual Figure

It is of interest to consider the dual figure obtained by subjecting the transducer diagram to the plane reciprocal transformation, studied in projective geometry, in which points transform into lines and lines into points. The two sets of collinear numbered points representing the input and output reactances become two concurrent sets of numbered lines. The point P becomes a line p.

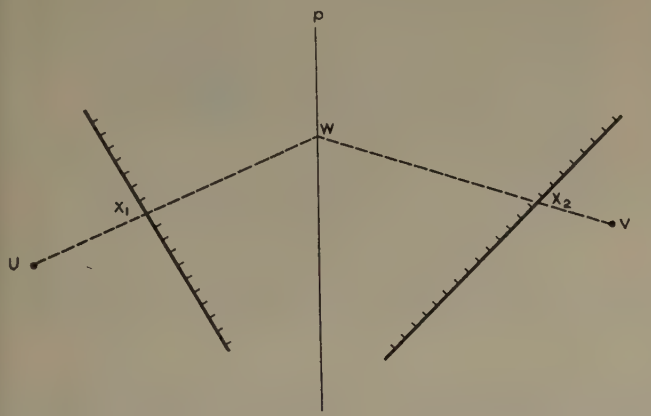

Fig. 5—The dual-transducer diagram obtained by reciprocation.

In Fig. 5 two graduated scales X_1 and X_2 serve to number the two pencils of lines through U and V. Given X_2, it is necessary to produce VX_2 to its intersection with p at W, and then to read off the value of X_1 corresponding to the line UW. This is the diagram given by Weissfloch.

Discussion

From the point of view of graphical treatment of physical problems, projective constructions are particularly simple. This holds both when accurate constructions are made for numerical purposes and also when rough sketches or mental pictures are used for qualitative thinking.

The bilinear function of a complex variable may be found running through the whole of communications theory because of the property of linear networks, that one and only one terminating impedance can give rise to a given input impedance, and conversely. The (1, 1) correspondence property is characteristic of linear networks, and the bilinear function gives the most general relation between complex variables in (1, 1) correspondence. But, under special conditions, the bilinear transformation of a complex variable can degenerate into a projective transformation of a real variable. Therefore, throughout the field where circle diagram techniques are used, there is the possibility of applying projective constructions.

In the present paper, the fact that the reactance axes correspond enables a projective construction to be applied, but it can be proved that in any transformation in which circles go into circles, there are always two straight lines (and two only) which go into straight lines.[3] Each of these is in projective correspondence with its transform. Extension of the present ideas to pairs of lines of this kind in the various fields of use of circle diagrams might well lead to some useful developments.

[3] Consider a particular dissipative transducer. Let the circle Q on the Z_1 plane correspond to the line $R_2=0$ on the Z_2 plane and let M be the point on Q corresponding to $Z_2=\infty$. Then the tangent to, and diameter of, Q at M correspond respectively to straight lines $R_2=$ const. and $X_2=$const. on the Z_2 plane.

A Method of Making a Radar Self-Calibrating*

E. A. WOLFF†, STUDENT MEMBER, IRE

Summary—**A method of making a radar system self-calibrating has been developed to obtain the ratio of power received to power transmitted with minimum effort. A portion of the transmitter pulse is made to bounce back and forth in a long, short-circuited, microwave-transmission line where it is delayed and attenuated. A portion of the power in the pulse is fed into the receiver at regular intervals to calibrate its output in terms of the transmitter power. Analytical expressions required for an understanding of the system have been developed. Several practical microwave-transmission lines have been studied. A self-calibrating system has been constructed, and the experimental results agree with theory.**

Introduction

ONE IMPORTANT APPLICATION of radar is the quantitative measurement of the radar area of targets. Radar area, σ, is defined as 4π times the ratio of the reflected power per unit solid angle in the direction of the source divided by the power per unit area in the incident wave. Thus, the determination of the radar area of a target involves the determination of the ratio of the power received at the receiver terminals, P_r, to the transmitter power, P_t. The radar area may be expressed as

$$\sigma = \frac{64\pi^3 R^4}{G^2\lambda^2}\,\frac{P_r}{P_t} = K\frac{P_r}{P_t}, \qquad (1)$$

* Decimal classification: R537.3XR537.4. Original manuscript received by the IRE, May 6, 1954; revised manuscript received, June 16, 1954. Work performed in partial fulfillment of the requirements for the Master of Science Degree at the University of Maryland.

† Naval Research Laboratory, Washington, D. C.

where G is the gain of the radar antenna, λ is the free-space wavelength, and R is the distance between radar and target.

In order to use a radar for the precise measurement of the radar area of a target, it is necessary to calibrate accurately the radar receiver output in terms of the ratio P_r/P_t. Since it is difficult to measure absolute power accurately, calibration procedures should depend only upon relative values of power. Several methods have been used with varying degrees of success to obtain an accurate, reliable calibration. These involve the careful measurement of the losses in each radar component, the use of "standard" targets, and the use of signal generators and power meters.

An echo box is sometimes used to give an indication of a radar's performance. It consists of a high Q resonant cavity which "rings" when excited by a portion of the transmitter pulse. The ringing produces an exponentially decaying signal whose duration is an indication of the over-all system performance. Since this signal does not have discrete steps which can be used to determine the ratio P_r/P_t, and it is difficult to keep the cavity properly tuned, the echo box is not suitable for accurately calibrating a radar in terms of the ratio P_r/P_t.

In one calibration method now in use, successive pulse levels, separated in power magnitude by a few decibels, are fed from a standard signal generator, through a directional coupler, into the radar receiver, and recorded.[1] The signal generator power at a particular attenuator setting is compared with the transmitter power by measuring both the transmitter and signal generator power with a common power meter. This process, along with the procedure to determine the ratio of the incident power at the antenna to the power at the receiver terminals, allows the construction of a plot of the receiver output vs σ. Relating the power at the receiver to the incident power at the antenna is a separate procedure which involves the measurement of the losses in the transmission lines and the insertion loss of the directional coupler.[2]

The calibration method now in use has four disadvantages. First, it consumes considerable time. Secondly, it requires careful tuning of the signal generator to the transmitter frequency (which may drift). Thirdly, the calibration pulses from the signal generator usually do not have the same spectrum or shape as the transmitted pulses. In the fourth place, it requires manual or machine switching of microwave transmission lines with the accompanying possibility of error. It was the attempt to devise a calibration procedure without these disadvantages that led to the development of the "self-calibration" system to be described.

SELF-CALIBRATION METHOD

In the self-calibration method, a portion of the transmitter pulse is made to bounce back and forth in a long microwave transmission line which has a short-circuit at each end. Each time the pulse reaches the input to the line, a portion of the pulse power is tapped off and fed into the receiver (Fig. 1). As the pulse travels along the line it is delayed and attenuated, so that the receiver picks up a succession of pulse-levels which form a calibration curve. The attenuation and delay will now be considered.

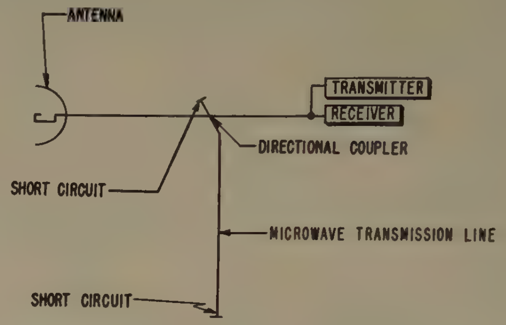

Fig. 1—Block diagram showing location of short-circuited transmission line.

With a rectangular radar pulse of duration t_p, the space distance between the leading and trailing edges is

$$d = vt_p, \tag{2}$$

where v is the velocity of propagation. Since the pulse travels up the transmission line and back (a distance equal to two line lengths) between steps, the minimum line length, $L_{\min}$, to separate two successive pulses is

$$L_{\min} = \frac{d}{2} = \frac{vt_p}{2}, \tag{3}$$

and the maximum attenuation per unit length of line, α, to produce an attenuation a_0 between steps is

$$\alpha = \frac{a_0}{d} = \frac{a_0}{vt_p}. \tag{4}$$

The total attenuation per step (α_0) will consist of the sum of the attenuations due to the line (a_0), the directional coupler which feeds the line (a_{dc}), and the discontinuities in the line (a_r). The attenuation of the line may be found from (4). The attenuations due to the directional coupler and to the discontinuities will now be considered.

Since a portion of the pulse bouncing back and forth in the line is tapped off each time the pulse passes through the directional coupler, there will be an attenuation of the pulse in the line due to the directional coupler of amount

$$a_{dc} = 20 \log \left[\frac{\text{antilog } (C/10)}{\text{antilog } (C/10) - 1} \right] \text{db}, \tag{5}$$

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[1] M. Katzin, "Quantitative radar measurements," PROC. I.R.E., vol. 35, pp. 1333–1334; November, 1947.

[2] *Ibid., loc. cit.*

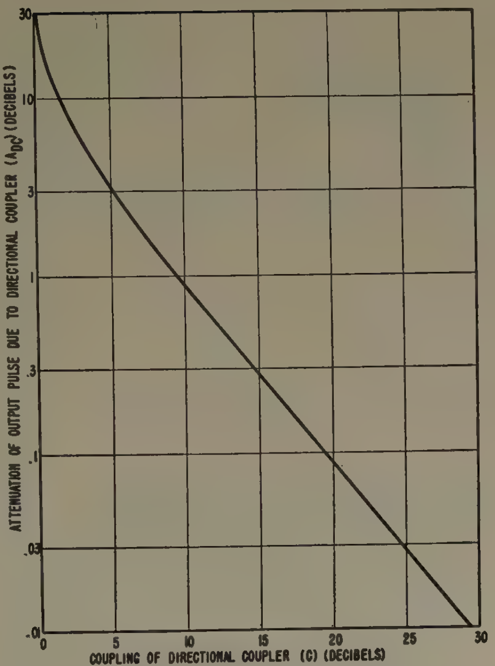

Fig. 2—Curve of attenuation of each output pulse as a function of directional coupler coupling.

where C is the coupling factor of the directional coupler in decibels (db). This relationship is plotted in Fig. 2.

If the transmission line contains a discontinuity, some of the power in the incident wave (P_{in}) will be reflected, and the transmitted power will be attenuated by an amount[3]

$$a_r = 10 \log \left(\frac{P_{\text{in}}}{P_t} \right) = 10 \log \left[\frac{(1+n)^2}{4n} \right] \text{db}, \tag{6}$$

where n is the voltage standing-wave-ratio caused by the discontinuity. Since there is a short-circuit at each end, the line may be considered to be infinitely long with an infinite number of directional couplers placed along the line at regular intervals, as indicated in Fig. 3. The outputs of the directional couplers are in parallel at the receiver input. From Fig. 3 it can be seen that a discontinuity may be represented as an infinite number of discontinuities distributed along the line. The waves reflected from these discontinuities appear at the receiver input as spurious pulses. These spurious pulses generally will not coincide in time with the desired calibrating pulses and will lower the peaks and raise the valleys of the desired pulse train. In addition, standing waves produce regions along the line where the current is greater than in the absence of standing waves. This increases the conductor losses and raises the value of α by the factor $(1+n^2)/2n$.[4]

Since reflections distort the output pulse train in a complicated manner and raise the value of α, construction of a good system in the beginning is simpler than trying to predict the results of a poorly constructed system.

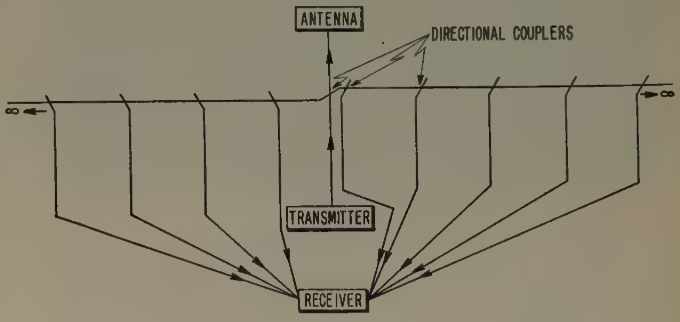

Fig. 3—Diagram showing how a transmission line with a short-circuit at each end may be represented as an infinite number of similar sections connected in series.

The total attenuation between steps, α_0, for a reflectionless line will be

$$\alpha_0 = a_0 + a_{dc}. \tag{7}$$

From (4), (5), and (7),

$$\alpha_0 = \alpha v t_p + 20 \log \left[\frac{\text{antilog } (C/10)}{\text{antilog } (C/10) - 1} \right] \text{db}. \tag{8}$$

It is desirable to have the calibration pulses arrive at the receiver after the recovery time so that the response of the receiver during the recovery time need not be determined. Since the pulses will be bouncing back and forth during the receiver recovery time, t_{rc}, an important consideration is the attenuation during the recovery time. The attenuation of the transmitter power during the recovery time, α_{rc}, will be the attenuation caused by the line during the recovery time, a_{rc}, and by two trips through the directional coupler. Therefore,

$$\alpha_{rc} = a_{rc} + 2C = \frac{t_{rc}\alpha_0}{t_p} + 2C \text{ db}. \tag{9}$$

If the width of the transmitter pulse is greater than the maximum pulse width for which the transmission line is designed (3), successive pulses will overlap and produce a resultant signal which will depend upon the relative magnitudes and phase angles of the overlapping pulses, and the shape of the pulses at the output may appear distorted.

The equations derived above provide the basis for the selection of practical transmission lines. As an aid to the calculation of parameters and the selection of transmission lines, a nomograph (Fig. 4, following page) has

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[3] G. L. Ragan, "Microwave Transmission Circuits," McGraw-Hill Book Co., Inc., New York, N. Y., p. 34; 1948.

[4] *Ibid.*, p. 31.

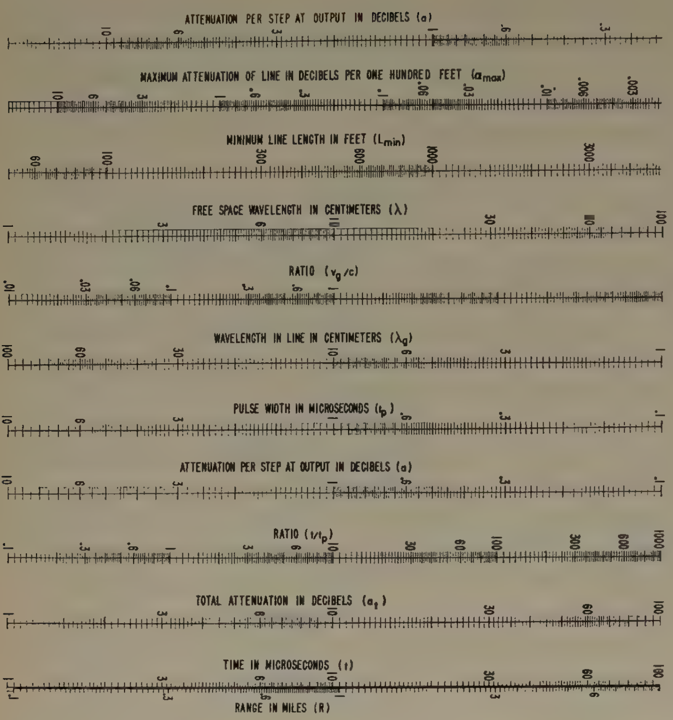

Fig. 4—Nomograph for selecting transmission lines.

Nomograph Instructions

To select a suitable transmission line when values of λ, t_p, a_0, and $L_{\min}$ are specified, the attenuation per unit length of line, α, and the group velocity, v_g, or waveguide wavelength, λ_g, are found by proceeding from column to column in the following order: 1-3-2-3-7-5-4-6.

To evaluate the performance of a line when values of λ, λ_g, or v/c, t_p and α are known, the attenuation at the output, a_0, and the minimum line length, $L_{\min}$, are found by proceeding from column to column in the following order: 4-6-5-7-3-2-1.

To evaluate the attenuation which occurs during any time interval when t_p and α_0 are known, proceed from column to column in the following order: 7-11-9-8-10.

been constructed. With the aid of the nomograph and published data on typical transmission lines, the line length, weight, and attenuation per step at the output for a one micro-second pulse were calculated and are given in Table I. This table has been used to aid in the determination of the stability of certain lines.

TABLE I

Characteristics of Practical Microwave Transmission Lines

Transmission Line	f MC/s	α DB/100′	λ_g cm.	$L_{\min}$ ft/μsec	α_0 DB/μsec	W_t lb/μsec
$1\frac{1}{2}''\times3''$ rect	3300	0.55	11.7	385	4.6	320
$\frac{5}{8}''\times1\frac{1}{4}''$ rect.	9375	2.5	3.86	410	24	112
$\frac{1}{2}''\times1''$ rect.	9375	3.4	4.48	352	26	57.4
3″ O.D. circ.	3300	0.5	12.8	352	3.6	238
1″ O.D. circ.	9375	3.0	5.17	310	19	35.1
RG-19A/U	1000	3.4	47	390	24	291
$\frac{3}{8}''$ Styroflex	1250	4.1	—	450	37	46
$\frac{3}{8}''$ Styroflex	3300	6.7	—	450	61	46
$\frac{1}{2}''$ Styroflex	1250	3.2	—	450	29	72
$\frac{7}{8}''$ Styroflex	1250	1.7	—	450	15	218
$1\frac{5}{8}''$ Styroflex	1250	0.97	—	450	9	410

The stability of lines in three frequency ranges has been studied. For this study only light, compact, commercial lines suitable for airborne use were considered. While circular waveguide, due to its low attenuation, might be superior to other lines, it was not studied because of practical difficulties prevailing at the time.

The variation of attenuation due to changes in frequency, temperature, and humidity have been investigated. The effect of the manufacturing tolerances on the theoretical attenuation has also been considered. The line considered for X band (9375 mc) was silver-plated rectangular copper waveguide ($1''\times\frac{1}{2}''$). The line for S band (3300 mc) was silver-plated rectangular copper waveguide ($1\frac{1}{2}''\times3''$). The line for L band (1250 mc) was $\frac{7}{8}''$ O.D. Styroflex[5] coaxial cable.

The attenuation in rectangular, silver-plated waveguide containing a lossless dielectric may be written as

$$\alpha = \frac{3.11\sqrt{\epsilon_r}\left[\frac{1}{b}+\frac{1}{2a}\left(\frac{\lambda}{a\sqrt{\epsilon_r}}\right)^2\right]}{\sqrt{\lambda}\sqrt{1-0.25\left(\frac{\lambda}{a\sqrt{\epsilon_r}}\right)^2}}\ \text{db/100 ft.,} \qquad (10)$$

where ϵ_r is the dielectric constant of the material in the waveguide, a and b are the large and small waveguide dimensions respectively (in centimeters), and λ is the free-space wavelength in centimeters. Using (10), the change in α for a 100 mc change in frequency for X-band waveguide is 0.03 db per 100 feet, and for S-band waveguide the change is 0.042 db per 100 feet.

The attenuation due to water vapor[6] (20 gm/cu. meter) within the waveguide at 73 degrees Fahrenheit is 0.00002 db per 100 feet at X band and less than this at S band.

Generally, the effects of normal changes in frequency, temperature, and humidity upon attenuation may be neglected for X-band and S-band waveguide.

Tabulated manufacturing tolerances substituted into (10) show that the maximum change in attenuation due to manufacturing tolerances for X-band waveguide is 0.04 db per 100 feet and for S-band waveguide is 0.02 db per 100 feet. Thus, the attenuation predicted by (10) using the nominal dimensions will be accurate within 2 per cent.

At L band (1250 mc), $\frac{7}{8}''$ O.D. Styroflex coaxial cable was investigated for stability. The attenuation due to conductor losses of a coaxial cable having different materials for the inner and outer conductors may be expressed as

$$\alpha_c = \frac{41.4\,(10)^3}{\lambda}\left[\frac{\delta_a\mu_a}{a}+\frac{\delta_b\mu_b}{b}\right]\frac{\sqrt{\epsilon_r}}{\ln\,(b/a)}\ \text{db/100 ft.} \qquad (11)$$

where a is the inner conductor outer diameter in cm, b, is the outer conductor inner diameter in cm, μ is the conductor permeability, δ is the conductor skin depth in cm, ϵ_r is the dielectric constant of the material within the cable,

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[5] Manufactured in the United States by Phelps Dodge Copper Products Corp.

[6] D. E. Kerr, "Propagation of Short Radio Waves," McGraw-Hill Book Co., Inc., New York, N. Y., pp. 660–661; 1951.

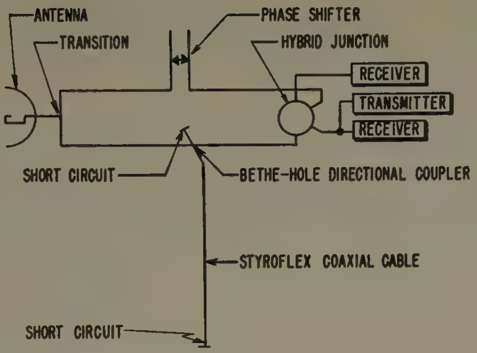

Fig. 5—Block diagram of microwave portion of a self-calibrating radar system.

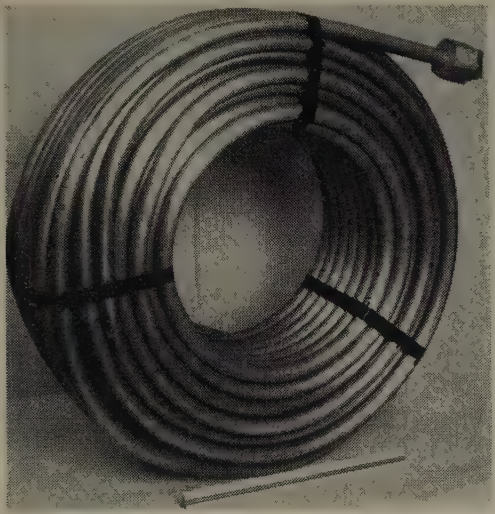

Fig. 6—Photograph of Styroflex transmission line.

and λ is the free space wavelength. The attenuation due to the dielectric losses in the cable may be expressed as

$$\alpha_d = \frac{8.32\ (10)^4\sqrt{\epsilon_r}\ \tan\gamma}{\lambda}\ \text{db/100 ft.,} \qquad (12)$$

where tan γ is the loss tangent of the dielectric. The effect of a change in frequency upon the value of α may be obtained from (11) and (12). For Styroflex, a 100 mc change in frequency changes the attenuation 0.042 db per 100 feet. The effects of temperature and humidity, small enough to be ignored for X-band waveguide, are even smaller for Styroflex and may also be ignored.

This study indicates that the three lines selected from Table 1 for a calibration system are sufficiently stable. In order to verify these calculations experimentally and to utilize this calibration method, a self-calibrating system was constructed. This will now be described.

Construction of Experimental Self-Calibrating System

The first experimental line and its requirements were selected after a study of Table I, the primary application of the radar, and the limitations imposed by airborne use. The calibrating system had to satisfy three requirements. First, it had to operate with pulse widths of at least $\frac{1}{2}$ microsecond. Secondly, the calibration curve had to contain at least eight "points" on order to yield the required accuracy. Thirdly, the system could not weigh more than 300 pounds for airborne use.

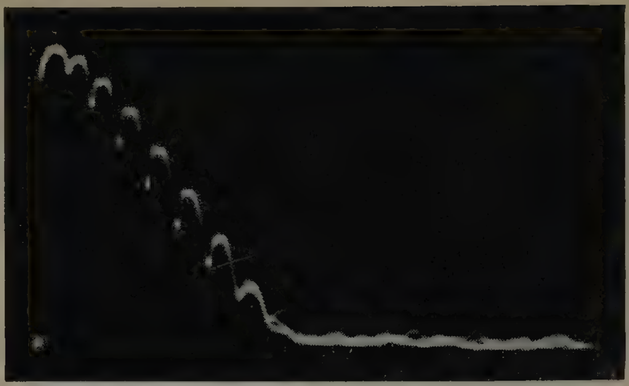

Fig. 7—Photograph of radar receiver A scope showing calibration pulses.

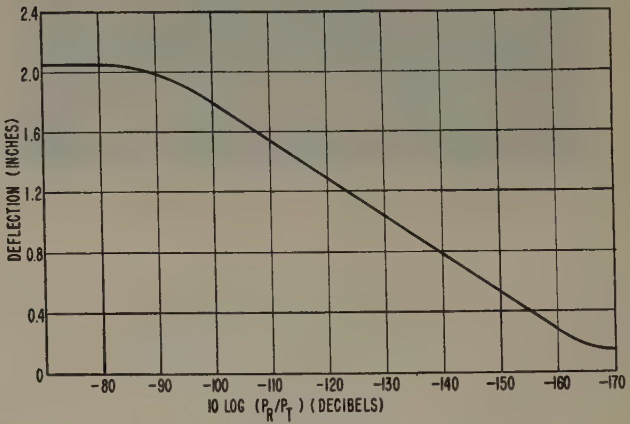

Fig. 8—Calibration curve for L band radar.

The line selected to meet these requirements for L band was $\frac{7}{8}''$ O.D. Styroflex. In the self-calibrating system, the Styroflex line replaces the signal generator and power meter. The microwave portion of the radar is illustrated in Fig. 5, with the Styroflex line added to show its relation to the other components of the system. The Styroflex line weighs 110 pounds and is 225 feet long, enabling the line to separate a $\frac{1}{2}$ microsecond pulse. The line was wound in a coil with an outer diameter of 23 inches and a length of 7 inches, Fig. 6.

The performance of the calibration system may be evaluated with the aid of Fig. 7, which is a photograph of an oscilloscope presentation of the receiver output (A scope), when the Styroflex is connected as shown in Fig. 5. The picture shows the large, saturated transmitter pulse followed by a succession of pulses (diminishing in magnitude) contributed by the Styroflex line. These pulses are separated in magnitude by 10.4 db and in time by $\frac{1}{2}$ microsecond. (The calculated value of attenuation for a reflectionless line was 7.7 db per step.) The final result, the calibration of the receiver output in terms of the transmitter power, is plotted in Fig. 8 from the curve in Fig. 7.

The performance of a line of half this length (which would separate a $\frac{1}{4}$ microsecond pulse) was determined by placing a directional coupler at each end of the Styroflex line, feeding power into one end, and taking power out at both ends of the line. In this way a portion of the pulse bouncing back and forth in the line is tapped off whenever the pulse arrives at either end of the line. Fig. 9 is a photograph of the receiver A-scope when using this arrangement. Because the directional couplers at each end of the line did not have exactly the same coupling, a large step in Fig. 9 is always followed by a smaller step. Thus, the steps at each directional coupler are equal, but the maximums of the pulses from one directional coupler do not appear half way between the pulses from the other directional coupler.

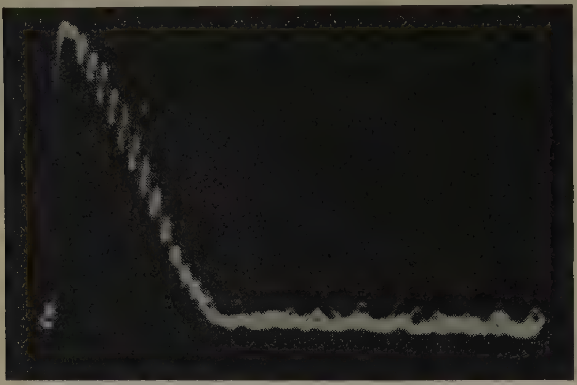

Fig. 9—Photograph of radar receiver A scope showing the effect of a shorter pulse and transmission line.

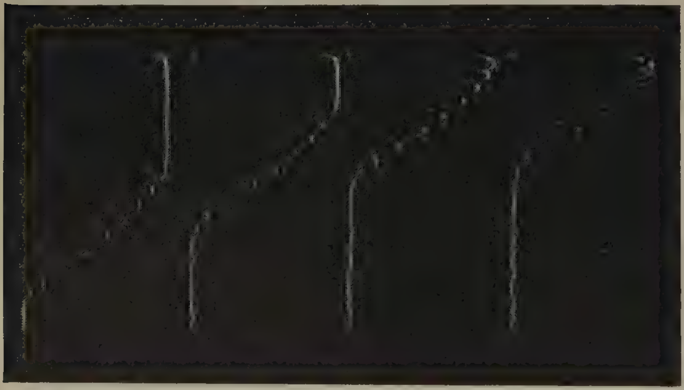

Fig. 10—Photograph of radar receiver A scope for transmitter pulse widths of $\frac{1}{4}$, $\frac{1}{2}$, $2\frac{1}{2}$, and 5 microseconds (top to bottom) and a $\frac{1}{2}$ microsecond transmission line.

Fig. 10 shows the results when pulse widths of $\frac{1}{4}$, $\frac{1}{2}$, $2\frac{1}{2}$, and 5 microseconds are used in the line designed to resolve a $\frac{1}{2}$ microsecond pulse. The distortion on the tops of the pulses, as described earlier, appears in all photographs except for the $\frac{1}{4}$ microsecond pulse. The distortion in the $\frac{1}{2}$ microsecond pulse is probably due to overlap of the tails of the neighboring pulses (caused by the lack of a perfectly rectangular transmitter pulse). Another interesting feature of Fig. 10 is the difference in slope of the echo decay between the various pulse lengths. Investigation showed that, as the pulse length was changed from $\frac{1}{4}$ to 5 microseconds the magnetron frequency was altered, so that the electrical length of the line was changed by more than a wavelength. Since the change of electrical length modifies the vector addition of the overlapping pulses, the absolute level of each pulse is shifted. A phase shifter may be used to compensate for the change in frequency but, due to the long length of line, it would be very frequency-sensitive. Thus, the system may be used for pulse widths greater than those for which the line was designed, but the results must be interpreted carefully.

When the Styroflex line was placed in a temperature-controlled room and its attenuation measured at -20 degrees F. and 120 degrees F. to an accuracy of ± 0.1 db, the change in attenuation was too small to be measured.

CONCLUSIONS

The features of the self-calibration method are:

1. It may be used to measure the radar input-output characteristic.
2. It may be used to provide a measure of the over-all radar system performance.
3. It is stable, as there are no active elements and no moving parts.
4. A calibration takes very little time.
5. It can be made relatively light in weight.

A comparison of Figs. 7 and 9 shows that a shorter line may be used to obtain more calibration points.

The accuracy of this method appears to be limited by the accuracy with which the A-scope deflection and transmission line attenuation can be measured. This should be $\pm\frac{1}{2}$ db or less.

The self-calibrating system should be useful in the measurement of the radar area of targets, and in the checking of the radar input-output characteristic. The system can be used to provide a frequent measure of the over-all radar performance of permanent installations (shipboard, airborne, landbased). The antenna can be included in the over-all radar performance check by placing the calibrating line, with its directional coupler and an antenna, at a reasonable distance from the radar.

ACKNOWLEDGMENT

The author is indebted to Mr. David L. Ringwalt, who suggested the calibration method described in this paper, and other members of the Wave Propagation Branch of the Naval Research Laboratory whose cooperation made possible the experimental part of the work.

A Series Laminated Conductor for High Frequencies*

HIKOSABURO ATAKA†

Summary—**To reduce the effective resistance due to skin effect in the conductor carrying the high frequency current, it is proposed that the conductor may be laminated and insulated at right angles to flow of the current, and it is called a series laminated conductor. In the series laminated conductor, it is shown that the distribution of the current becomes skin effect in pieces of the conductor, while bone effect in layers of the dielectric, and it is concluded that its effective resistance will become smaller than that of the solid conductor.**

INTRODUCTION

IF THE ATTENUATION of a coaxial cable for high frequencies is analyzed in parts, it will be found that more than fifty per cent of the attenuation is caused by the increase of the resistance of its inner conductor, due to skin effect. Many attempts are made to reduce the resistance, caused by this phenomenon. Particularly Clogston and his colleagues have proposed to laminate the conductor and to insert it in the proper dielectric medium.[1] In this case, however, the lamination is performed in parallel along the direction of the flow of current. For the sake of convenience, a conductor made of this type of lamination will, here, be called a parallel laminated conductor.

Contrary to this lamination, as explained below, a conductor may also be laminated at right angles to the direction of the flow of current. A conductor made with this type of lamination will be called a series laminated conductor.

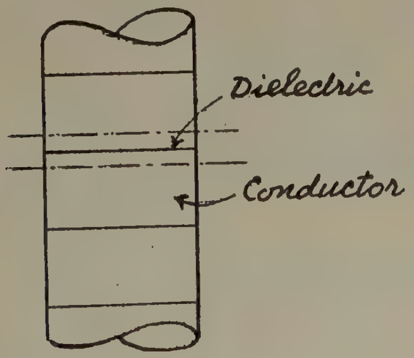

Fig. 1—A series laminated conductor.

A SERIES LAMINATED CONDUCTOR

When the high frequency current flows through the series laminated conductor as shown in Fig. 1, it is natural to suppose that the current will distribute under the control of the skin effect, in the middle part of the pieces of the conductor between the dielectric layers. But it is also natural to suppose that the current will distribute under the different law, in the neighborhood of two ends facing dielectric layers. In order to consider the point, draw out this part of conductor as shown in Fig. 2, and take the cylindrical shell of radius ρ, of thickness $d\rho$, and of length τ. Let

σ = the density of current at distant ρ,
H = the magnetic force at distant ρ,
ϕ = number of flux interlinked with the shell,
e = the impressed emf,
ω = the angular frequency of the emf,
$\delta\tau$ = the thickness of dielectric layer,
k = the conductivity,
ϵ = the dielectric constant,
μ = the permeability.

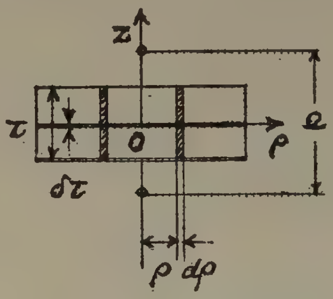

Fig. 2—Part of the conductor near the dielectric layer.

The resistance R and the capacitance C of the shell is given by:

$$R = \tau(1-\delta)/(k2\pi\rho d\rho), \qquad C = \epsilon 2\pi\rho d\rho/(\delta\tau); \tag{1}$$

and by Kirchhoff's law we have:

$$\sigma R 2\pi\rho d\rho + \frac{\partial\phi}{\partial t} + \frac{1}{C}\int_0^t \sigma 2\pi\rho d\rho dt = e. \tag{2}$$

Substituting (1) in (2) and differentiating with respect to ρ, we have:

$$(1-\delta)\frac{\partial\sigma}{\partial\rho} + \frac{k}{\tau}\frac{\partial^2\phi}{\partial\rho\partial t} + \frac{\delta k}{\epsilon}\int_0^t \frac{\partial\sigma}{\partial\rho}dt = \frac{k}{\tau}\frac{\partial e}{\partial\rho}. \tag{3}$$

In the conductors facing on the thin dielectric, we may put $\partial e/\partial\rho$. Thus,

$$(1-\delta)\frac{\partial\sigma}{\partial\rho} + \frac{k}{\tau}\frac{\partial^2\phi}{\partial\rho\partial t} + \frac{\delta k}{\epsilon}\int_0^t \frac{\partial\sigma}{\partial\rho}dt = 0. \tag{4}$$

Next the relation between σ and ϕ is given by Ampere's law,

$$\phi = \int_\rho^\infty \mu H\tau dr, \qquad \therefore \quad \frac{\partial\phi}{\partial\rho} = -\mu H\tau; \tag{5}$$

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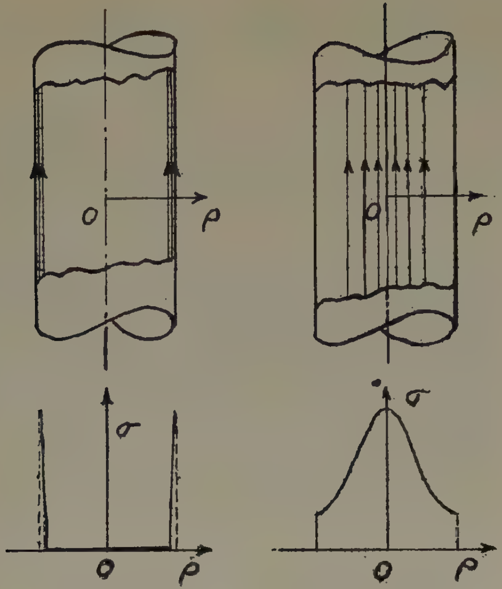
* Decimal classification: R117.2×R144.2. Original manuscript received by the IRE, February 2, 1954; revised manuscript received, May 18, 1954.

† Ibaraki University, Japan.

[1] A. M. Clogston, "Reduction of skin effect losses by the use of laminated conductors," *Bell Sys. Tech. Jour.*, vol. 30, pp. 491–529; July, 1951.



Conductor Dielectric

Fig. 3—Distributions of the density of current and of the lines of flow.

and,

$$2\pi\rho H = \int_0^\rho \sigma 2\pi r dr, \qquad \therefore \quad H = \frac{1}{\rho}\int_0^\rho \sigma r dr. \tag{6}$$

From (5) and (6) we get:

$$\frac{\partial\phi}{\partial\rho} = -\frac{\mu\tau}{\rho}\int_0^\rho \sigma r dr. \tag{7}$$

Under the sinusoidal emf e, both σ and ϕ become also sinusoidal, so it may be put:

$$\frac{\partial\phi}{\partial t} = i\omega\phi, \qquad \int_0^t \sigma dt = \frac{\sigma}{i\omega}, \qquad (i = \sqrt{-1}). \tag{8}$$

By (4), (7), and (8), we get finally:

$$\frac{d^2\sigma}{d\rho^2} + \frac{1}{\rho}\frac{d\sigma}{d\rho} - \frac{i\omega\mu k}{(1-\delta) - i(k\delta/\omega\epsilon)}\sigma = 0, \tag{9}$$

which is Bessel's equation of the zeroth order.

Solving (9) we have:

$$\sigma = \sigma_a \frac{J_0(\gamma\rho)}{J_0(\gamma a)}, \qquad (0 < \rho < a), \tag{10}$$

where a is the radius of the conductor, and σ_a is the density of current at $\rho = a$, and

$$\gamma = \sqrt{\frac{-i\omega\mu k}{(1-\delta) - i(k\delta/\omega\epsilon)}}. \tag{11}$$

Now, if we put $\delta = 0$, then we have:

$$\gamma = \sqrt{-i\omega\mu k}, \qquad \sigma = \sigma_a \frac{J_0(\gamma\rho)}{J_0(\gamma a)}.$$

In this case, the conductor is continuous without being separated by dielectric layers and the distribution of the current becomes skin effect. On the contrary, if we put $\delta = 1$, then we have:

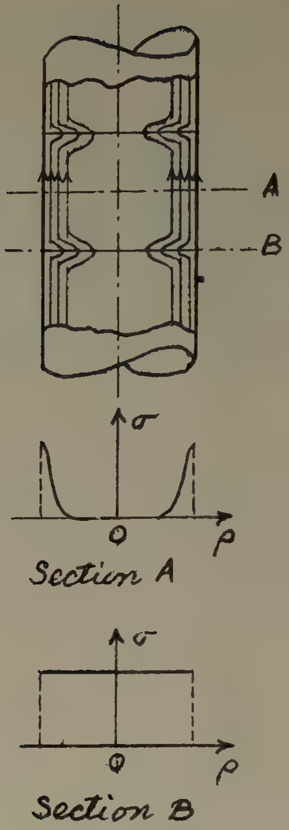


Fig. 4—Distribution of the density of current at Sections A and B, and distributions of the lines of flow of the series laminated conductor.

$$\gamma = \omega\sqrt{\epsilon\mu}, \qquad \sigma = \sigma_a \frac{J_0(\gamma\rho)}{J_0(\gamma a)}.$$

In this case, the conductor is replaced wholly by dielectric layer and the distribution of the current becomes bone effect.[2]

Thus, when the high frequency current flows through a cylindrical medium, the density σ and the lines of flow of the current will distribute as shown in Fig. 3, according to whether the medium is the conductor or the dielectric.

In the neighborhood of two ends facing the dielectric layer of the series laminated conductor, it may be put that,

$$1 - i(k\delta/\omega\epsilon) \doteqdot -i(k\delta/\omega\epsilon), \qquad \therefore \quad \gamma \doteqdot \omega\sqrt{\frac{\epsilon\mu}{\delta}}.$$

To get the concrete idea, some numerical values are put in these results. Let the conductor be of copper of the radius 1 mm, and the dielectric be a material of $\epsilon_s = 2$ and of the thickness $\delta = 10^{-4}$. The frequency of the impressed emf is taken to be 100 mc.

$$k = 5.70 \times 10^{-7}\ \mho/m, \qquad \epsilon = 2 \times 8.86 \times 10^{-12}\ F/m,$$
$$\mu = 1.26 \times 10^{-6}\ H/m, \qquad \omega = 2\pi \times 10^{8}\ C/s,$$
$$\delta = 10^{-4}, \qquad a = 10^{-3}\ m.$$

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[2] H. Ataka, "On the properties of an isotropic dielectric in the electric field of ultra high frequencies," *Jour. IEE of Japan*, no. 620, p. 116; 1940.

In this case,

$$k\delta/\omega\epsilon = 5.12 \times 10^5, \qquad 1 - i(k\delta/\omega\epsilon) \doteqdot - i(k\delta/\omega\epsilon).$$

$$\gamma = \omega\sqrt{\frac{\epsilon\mu}{\delta}} = 2.97 \times 10^2\ m^{-1}.$$

$$\therefore \quad \sigma = \sigma_a \frac{J_0(29.7\rho)}{J_0(0.297)} \doteqdot \sigma_a, \qquad (0 < \rho < 10^{-3}).$$

Thus, in the series laminated conductor, the distributions of current and of the lines of flow will become as shown in Fig. 4. By inserting thin dielectrics, the skin depth in the pieces of conductor will become greater than that of the solid conductor.

CONCLUSION

The skin effect in the solid conductor is compensated by the bone effect, by cutting the conductor into pieces and separating with thin dielectric layers. The compensation becomes clearly the more effective, the shorter the length of individual pieces. If the conductor is cut indefinitely short and separated with thin dielectric layers, then it can be proven that the current distributes almost uniformly throughout the section. From the practical standpoint, however, there is an optimum length for individual pieces. The determination of this length will need more elaborate analysis.

Analysis of Transmission Line Directional Couplers*

W. L. FIRESTONE†, SENIOR MEMBER, IRE

Summary—**Five different types of transmission line systems are analyzed and are shown to possess directional properties at the lower radio frequencies. The conditions of mismatch on either transmission line or on both transmission lines have been taken into account, as well as lengths of coupling greater than a fractional wavelength. A new and more general scatter matrix for directional couplers which includes the condition of mismatch is developed. In addition, lumped circuit directional couplers have been developed as a consequence of the general transmission line theory.**

INTRODUCTION

THE USE of directional couplers as a means of measuring reflections, monitoring power, mixing, and multiplexing has been well known for many years.[1] Although many different types of waveguide-directional couplers are now in use, the purpose of this paper is to analyze the directional coupling effects between open-wire transmission lines which are used at the lower radio frequencies.

Due to the wide use of directional couplers, several definitions have come into being. However, the author wishes to propose the following definition, which will subsequently be justified.

A directional coupler is defined as a junction of four transmission lines (1), (2), (3), and (4), terminated in such a way that the normalized impedance products $Z_{11}Z_{44}$ equal unity and $Z_{22}Z_{33}$ equal unity, and there is no coupling between (1) and (4), and (2) and (3), but coupling does exist between the other terminals.

The above definition conforms to the text material but differs from the standard definition in that terminals (3) and (4) are reversed. It is to be noted that this definition does not carry the restriction of matched lines.

Although many excellent papers[2–9] on directional coupling have been published, relatively little work has been done with open wire lines[10–12] operating at the lower radio frequencies.

BASIC RELATIONS FOR A FOUR WIRE SYSTEM

The author has previously shown[13,14] that a matched four-wire transmission line system will display directional coupling characteristics if the following relationship is maintained,

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* Decimal classification: R117. Original manuscript received by the IRE, June 26, 1953; revised manuscript received, June 14, 1954.

† Motorola, Inc., Chicago, Ill.

[1] C. G. Montgomery, "Principles of Microwave Circuits," Lab. Series, McGraw-Hill Inc., New York, N. Y., vol. 8, pp. 148, 300–302; 1948.

[2] H. J. Riblett, "A mathematical theory of directional couplers," Proc. I.R.E., vol. 35, p. 1307; November, 1947.

[3] Andrew Alford, "Coupled networks in radio frequency circuits," Proc. I.R.E., vol. 29, pp. 55–70; February, 1941.

[4] N. I. Korman, "Theory and Design of Several Types of Wave Selectors," National Electronics Conference; October 4, 1946.

[5] W. W. Mumford, "Directional couplers," Proc. I.R.E., vol. 35, pp. 160–166; February, 1947.

[6] S. Rosen and J. T. Banger, "A consideration of directivity in waveguide directional couplers," Proc. I.R.E., vol. 37, pp. 393–402; April, 1949.

[7] E. L. Ginzton and P. S. Goodwin, "A note on coaxial Bethe hole directional couplers," Proc. I.R.E., vol. 38, pp. 305–310; March, 1950.

[8] H. C. Early, "A wide band directional coupler for waveguide," Proc. I.R.E., vol. 34, pp. 883–886; November, 1946.

[9] H. J. Riblett and T. S. Saad, "A new type of waveguide directional coupler," Proc. I.R.E., vol. 36, pp. 61–64, January, 1948.

[10] R. H. Miller, "A proposal for broad band coupling oc power line carrier equipment," *Trans. A.I.E.E.*, vol. 68, pt. 2, pp. 1028–1031; 1949.

[11] H. J. Sutton, "Line tuning equipment used with coaxial cable for carrier current installation on power lines," *Trans. A.I.E.E.*, vol. 68, pt. 1, pp. 44–49; 1949.

[12] E. Hancess, "Carrier current communications over high voltage transmission lines," *Brown Boveri Review*, vol. 31, 10, pp. 335–339; October, 1945.

[13] W. L. Firestone, "Transmission Line Directional Couplers," Doctoral Thesis, Northwestern University, May, 1952.

[14] W. L. Firestone, "Directional Coupling with Transmission Lines," National Electronics Conference; September 28, 1953 and *Tele. Tech.*, p. 95; October, 1953.

$$\frac{M}{C_m} = Z_{01}Z_{02} \text{ (characteristic equation)}, \qquad (1)$$

where M is the mutual inductance per unit length between the two coupled transmission lines, C_m is the mutual capacity per unit length, and Z_{01} and Z_{02} are the respective characteristic impedances. It was also shown that

$$C_m = \frac{4\epsilon_r(10^{-7})}{Z_{01}Z_{02}} \ln \sqrt{\frac{D_{23}D_{14}}{D_{13}D_{24}}},$$

and

$$M = 4(10^{-7}) \ln \sqrt{\frac{D_{23}D_{14}}{D_{13}D_{24}}} \cdot \qquad (2)$$

Furthermore, the coupling equation for such a system was given by

$$Db_c = 20 \log_{10} \left\{ \frac{4wl(10^{-7}) \ln_\epsilon}{\sqrt{Z_{01}Z_{03}}} \frac{\sqrt{\frac{D_{14}D_{23}}{D_{13}D_{24}}}}{} \right\} l < 0.1\lambda. \qquad (3)$$

MISMATCHED TRANSMISSION LINES

Directivity

Consider the arrangement shown in Fig. 1 in which Z_{L3} and Z_{14} are connected to terminals which are at some distance from the ends of the coupled section.

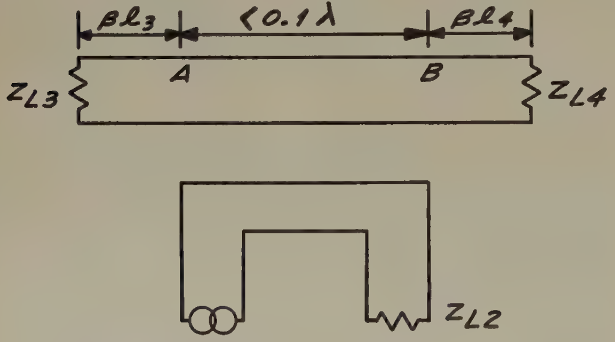

Fig. 1—Configuration for reflections on both lines.

It is desirable to refer all terminations to the ends of the coupling element itself. With the help of the standard transmission-line equations, Z_{L3} is referred to point A and Z_{L4} is referred to point B. Thus it is possible to restrict the discussion to a coupling circuit where everything is in phase.

With the aid of the following relationships

$$e_3 = -M\frac{di}{dt}, \qquad \frac{dq_{3C}}{dt} = -C_m\frac{dv_{12}}{dt} \qquad (4)$$

and Kirchoff's laws, it follows that

$$\text{(a)} \qquad -M\frac{di_1}{dt} - (Z_{L3}' + Z_{L4}')i_{3M} = 0. \qquad (5)$$

Eqs. (4) and (5) can be solved, respectively, for the inductive and capacitive currents i_{3M} and i_{3C}, which flow in the secondary line, yielding

$$i_{3M} = \frac{-MV_0}{(Z_{L3} + Z_{L4})Z_{L2}'} j\omega\epsilon^{j\omega t} \qquad (6)$$

and

$$i_{3C} = -\frac{C_m}{2} V_0 j\omega\epsilon^{j\omega t}, \qquad (7)$$

since

$$V_{12} = V_0\epsilon^{j\omega t} \quad \text{and} \quad I_0 = V_0 Z_{12}. \qquad (8)$$

Since the two end sections act in a parallel sense in supplying induced charge, the amount supplied from each end will depend upon the relative impedance at points A and B. Hence, the amount of capacitive current flowing to the left at point A may be expressed as

$$i_{3ce} = -j\omega C_m\left(\frac{Z_{L4}'}{Z_{L3}' + Z_{L4}'}\right) V_0\epsilon^{j\omega t}, \qquad (9)$$

while that flowing to the right at point B may be expressed as

$$i_{3cr} = +j\omega C_m\left(\frac{Z_{L3}'}{Z_{L3}' + Z_{L4}'}\right) V_0\epsilon^{j\omega t}. \qquad (10)$$

The total current to the left (i_{3e}) is the sum of (6) and (9) while the total current to the right (i_{3r}) may be expressed as the sum of (6) and (10).

Since the powers delivered to junctions 3 and 4 are due to the magnitude of the respective currents squared times the real part of the terminating impedance, we may write after some algebraic manipulation the following directivity ratio.

$$\frac{P_{L3}}{P_{L4}} \left| \frac{1 + Z_{22}Z_{44}'}{1 - Z_{22}Z_{44}'} \right|^2 \frac{R_{33}'}{R_{44}'} \qquad (11)$$

where all the Z's and R's are normalized with respect to the proper characteristic impedance.

The expression of (11) in decibels yields

$$DB_d = 20 \log \left\{ \sqrt{\frac{R_{33}'}{R_{44}'}} \left| \frac{Z_{22}Z_{44}' + 1}{Z_{22}Z_{33}' - 1} \right| \right\} . \qquad (12)$$

In the event that $\beta l_3 = \beta l_4 = 0$ (as shown in Fig. 1) and Z_{L3} and Z_{L4} and Z_{L2} are purely resistive, (12) becomes

$$DB_d = 20 \log \left\{ \sqrt{\frac{Z_{33}}{Z_{44}}} \left| \frac{Z_{22}Z_{44}' + 1}{Z_{22}Z_{33}' - 1} \right| \right\} \begin{array}{l}\text{All } Z\text{'s} \\ \text{Resistive.}\end{array} \qquad (13)$$

Fig. 2 is a plot of (13) when only Z_{33}' is mismatched and caused to vary. The experimental points corroborate the theoretical conclusions.

A little consideration of (12) will show that, if $Z_{22}Z_{33}=1$, an infinite directional coupler will result regardless of the other terminations. This is a surprising result and means that any degree of mismatch at terminals (2) or (3) is tolerable, providing Z_{22} and Z_{33}' are reciprocals. In other words, the condition that $Z_{33}=Z_{33}'=1$, or that matched conditions must be achieved, is strictly a special case of $Z_{22}Z_{33}'=1$. If the input terminals are considered to be at terminals (4), then we can readily see that, if $Z_{22}Z_{33}'=1$, an infinite directional coupler will still obtain.

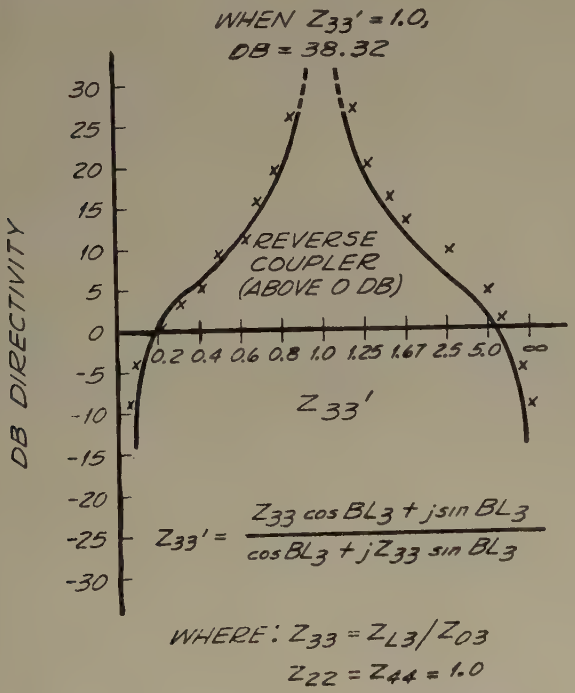

Fig. 2—Db directivity vs normalized impedance.

It is now easy to see that, if either terminals (2) or (3) are considered as input terminals, the necessary condition for infinite directional coupling to obtain is that $Z_{11}Z_{44}'=1$. Fig. 3 adequately substantiates the preceding theory of reciprocal impedances. For Figs. 2 and 3, $Z_{mm}'=Z_{mm}$. The possibility of having a directional coupler with infinite directivity, even though all the terminals are mismatched, has great practical applications.

If in (13), all the terminations except Z_{22} are matched (i.e.: $Z_{33}=Z_{44}=1$) we obtain:

$$DB_d = 20 \log \left| \frac{Z_{22}+1}{Z_{22}-1} \right|. \tag{14}$$

It is interesting and informative at this point to note that, if Z_{22} is replaced with its reciprocal $(Z_{22})^{-1}$, the directivity is unchanged. Fig. 4 is a plot of (14). The added experimental points verify the basic theory of reflections on the main line.

For the coupling ratio we may write

$$\frac{P_{L3}}{P_{11}} = \frac{|\, i_{3l} \,|^2 R_{L3}'}{|\, V_{12} \,|^2 / R_{L2}}. \tag{15}$$

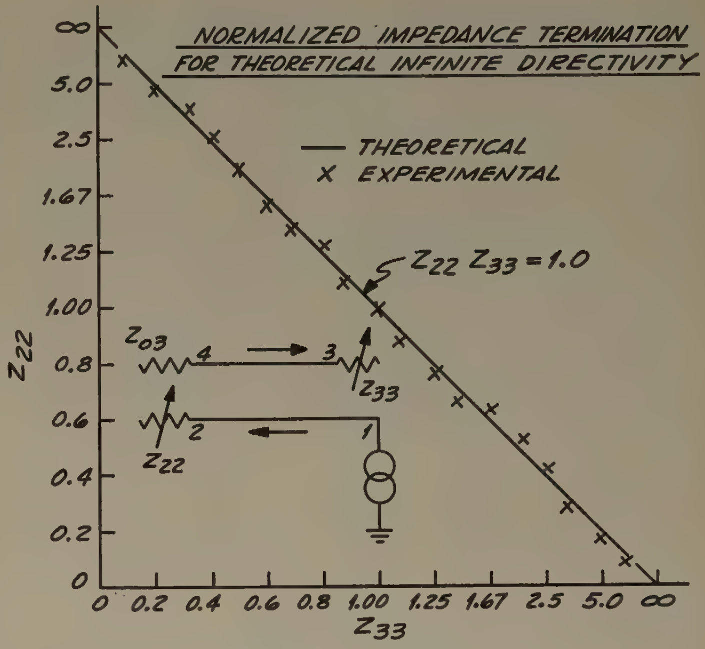

Fig. 3—Normalized impedance termination for theoretical infinite directivity.

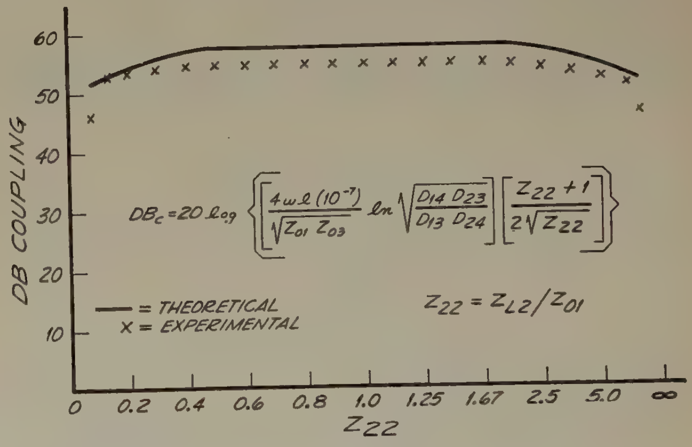

Fig. 4—Db directivity vs normalized impedance.

By use of the proper expression for i_{3e}^2, and by using rms values, we can obtain the coupling equation expressed in decibels as

$$DB_c = 20 \log \left\{ \left[\frac{4wl(10^{-7})}{\sqrt{Z_{01}Z_{03}}} \ln \sqrt{\frac{D_{14}D_{23}}{D_{13}D_{24}}} \right] \cdot \left| \frac{\sqrt{R_{22}R_{33}}}{Z_{33}'+Z_{44}'} \left(\frac{1}{Z_{22}} + Z_{44}' \right) \right| \right\}. \tag{16}$$

We now observe that the part of the log term which is in the first rectangular bracket is identical to the case where no reflections are present. (See (3).) We therefore see that the effect of reflections on the two lines is to add this second bracketed term. Furthermore, if $Z_{22}=Z_{33}'=Z_{44}=1$ (no reflections), this term reduces to unity, which is exactly what we expect to happen. We may call this our "general loss factor." If $Z_{33}'=Z_{44}'=1$, the above factor reduces to

$$\left|\frac{R_{22}+1}{2\sqrt{R_{22}}}\right|$$

which we may call simply the "loss factor." The "loss factor" is obviously to be related to coupled transmission lines where reflections exist only on the main line. This loss factor has the characteristic that if Z_{22} is replaced by its reciprocal, the term remains the same. Fig. 5 shows the theoretical curve, as well as experimental points which were taken as supporting evidence for the coupling equation when Z_{22} is varied.

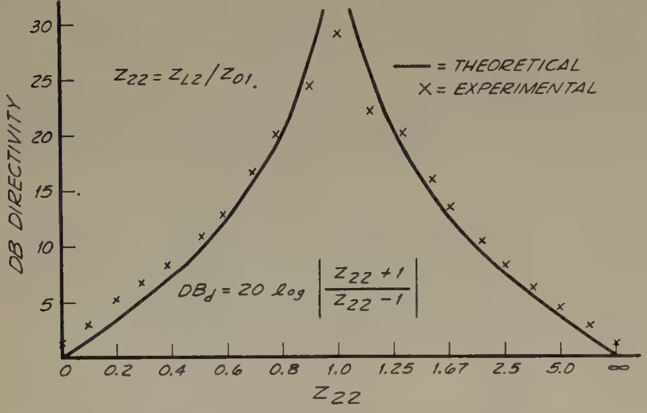

Fig. 5—Length of coupling.

For $Z_{22}=Z_{44}'=1$, the General Loss Factor becomes

$$\left|\frac{2\sqrt{R_{33}'}}{1+Z_{33}'}\right|.$$

In the preceding section we observed the interesting and valuable condition that an infinite directional coupler is obtained whenever $Z_{22}Z_{33}'=1$. It appears desirable to see what happens to the coupling whenever this condition obtains.

The general loss factor mentioned heretofore reduces to a constant, namely unity, when $Z_{22}Z_{33}'=1$. This means that the coupling remains a constant as, say, Z_{33}' is varied, providing the reciprocal relationship between Z_{22} and Z_{33}' is maintained. Because of space limitations, corroborating evidence of the above conclusions will not be presented here.[2]

The total power coupled to the secondary line in either direction can readily be found to be

$$DB_c = 20\log\left\{\left[\frac{4wl(10^{-7})}{\sqrt{Z_{01}Z_{03}}}\ln\sqrt{\frac{D_{14}D_{23}}{D_{13}D_{24}}}\right]\right.$$

$$\left.\cdot\sqrt{\frac{Z_{22}{}^2+1}{2Z_{22}}}\right\}. \tag{17}$$

SCATTERING MATRIX

From the above, it is apparent then that a mismatch at one pair of terminals can set up a reflection which is cancelled (insofar as directional coupling is concerned) by a mismatch at another pair. Besides being a useful and instructive phenomenon, these relations permit us to formulate a more general scattering matrix than has been done in the past. The new scattering matrix, derived in Appendix 1, is given below and includes the mismatched conditions.

$$S=\left[\begin{array}{cc|cc} \bar{\gamma} & j\beta_3 & 0 & j\beta_2 \\ j\beta_3 & \bar{\gamma}^* & j\beta_2 & 0 \\ \hline 0 & j\beta_2 & \bar{\gamma} & -j\beta_3 \\ j\beta_2 & 0 & -j\beta_3 & \bar{\gamma}^* \end{array}\right]. \tag{18}$$

The various elements of the matrix are scattering coefficients and are a measure of the amplitude of the wave scattered into one pair of terminals by an incident wave of unit amplitude on another pair of terminals. If we consider matched lines only, it can readily be shown[2] that the above matrix reduces to the one given below, which up to now had been regarded as the standard form for a directional coupler.

$$s=\left[\begin{array}{cc|cc} 0 & \alpha & 0 & j\beta \\ \alpha & 0 & j\beta & 0 \\ \hline 0 & j\beta & 0 & \alpha \\ j\beta & 0 & \alpha & 0 \end{array}\right]. \tag{19}$$

DIRECTIVITY AND COUPLING FOR ANY LENGTH OF COUPLED LINE

The purpose of this section is to determine the directivity and coupling between two transmission lines terminated in their characteristic impedances without the restriction that the coupled section be small compared to a wavelength.

At any point on the secondary line where coupling occurs, waves are induced due to the mutual inductance and mutual capacitance effects. The mutual inductance effect results in waves traveling both to the right and to the left on the secondary line. The mutual capacitance effect also results in waves traveling to the right and to the left on the secondary line. However, it has already been shown that a phase difference exists between the waves traveling at the right of the coupled section, which results in cancellation to the right. To the left of the coupled section, however, these waves add and consequently a resultant wave travels to the left. This is true for every point of coupling. However, the manner in which these individual waves add together at the left-hand termination will be a function of their relative phases. For a long coupling length, phase shifts will occur.

At any point along the main line, the voltage may be expressed as

$$V_{12}V_0\epsilon^{j\omega t}\epsilon^{-j\beta x} = K_2\epsilon^{-j\beta x}. \tag{20}$$

In each region along the main line from the point x, a resultant backward wave is induced on the secondary line. The voltage associated with this wave referred to $x=o$ (i.e.: generator pt.) may be represented by

$$de_3 = K_2V_{12}\epsilon^{-j\beta x}dx = K_3\epsilon^{-2j\beta x}dx, \tag{21}$$

where K_3 is the resultant voltage induced on the secondary line per unit length, and e_3 is the total induced voltage per meter of coupling length. If the coupling extends over a region from 0 to l, the voltage of the resulting wave on auxiliary line is given by above equation integrated from 0 to l. After integration, this yields

$$e_{3l} = K_3\epsilon^{-j\beta l}\left(\frac{\sin(\beta l)}{\beta}\right). \tag{22}$$

To obtain K_3, it is merely necessary to multiply the total current induced into line (3, 4), for a unit length of coupling, by Z_{03} and take the magnitude of the resultant equation. The total current induced into line (3, 4), per unit length of coupling is calculated as previously stated. It follows then that

$$|K_3| \doteq \frac{wI_0}{\sqrt{2}}4(10^{-7})\ln\sqrt{\frac{D_{14}D_{23}}{D_{13}D_{24}}}. \tag{23}$$

Since $P_{L3}=|e_3|^2/Z_{03}$, the coupling ratio for any length is

$$\frac{P_{L3}}{P_{11}} = \left|\frac{e_{3\epsilon}}{V_{12}}\right|^2\frac{Z_{01}}{Z_{02}}. \tag{24}$$

After substituting (20), (22), and (23) into (24) and expressing (24) in decibels, we obtain

$$DB_c = 20\log\left\{\left[\frac{4w(10^{-7})}{\sqrt{Z_{01}Z_{02}}}\ln\sqrt{\frac{D_{14}D_{23}}{D_{13}D_{24}}}\right]\cdot\left[\frac{\sin(\beta l)}{\beta}\right]\right\}. \tag{25}$$

We now observe that when βl is small, (23) reduces to (3), which is the expected result for small βl.

A plot of (22) and (22) squared is given in Fig. 5. Fig. 5 points to the feasibility of reversing the phase of coupling every quarter of a wavelength. Another possible arrangement that is even more effective in transferring power would be one in which 240-degree phase shifters are introduced at 60-degree phase intervals along the primary line.

OTHER CONFIGURATIONS

A. A Two Wire System with Ground Plane

One conductor of both the primary and the secondary lines may be replaced by a ground plane. For a cross section of such a configuration see Fig. 6. The ground plane is used as a return circuit for both of the transmission lines.

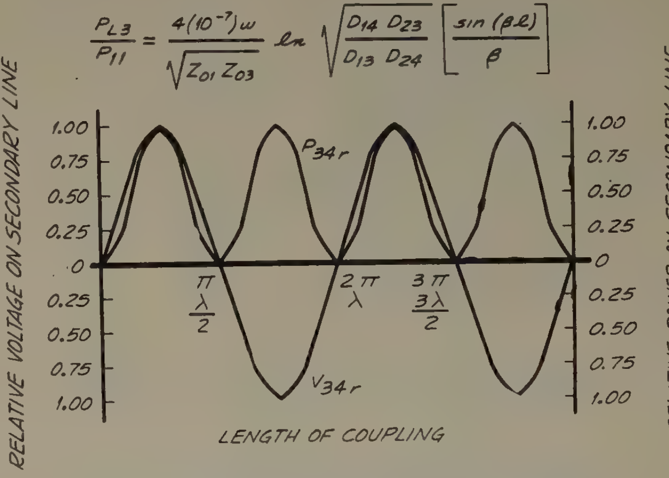

Fig. 6—Two single-wire transmission lines using a common ground plane.

The characteristic impedances of such transmission lines are given by[15]

$$Z_{03} = 138\log_{10}\frac{2h_3}{r_3},$$

and

$$Z_{01} = 138\log_{10}\frac{2h_2}{r_2}. \tag{26}$$

By the method of imaging, the two-wire system can be treated mathematically as a four-wire system as shown in Fig. 7 on page 1534.

It follows that the coupling and directivity equations are identical for the two systems (see (16) and (13)). It should be remembered that the values of the characteristic impedances to be used in making computations of coupling and directivity for the two-wire system are the same as those used in making the same computations for the comparable four-wire system.

Even though the coupling for the two-wire and four-wire systems is the same, the power in the one-wire lines is 3 db less than the power in the comparable two-wire lines.

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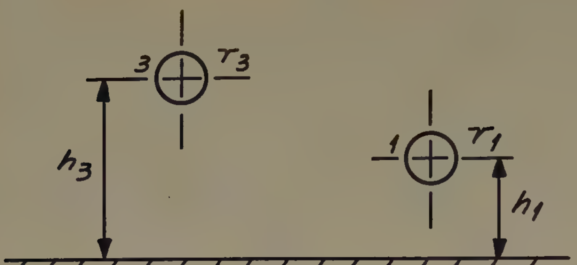
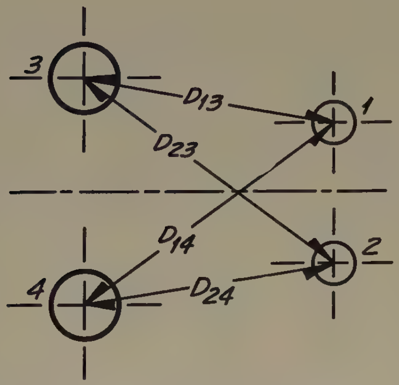
[15] C. P. Harnwell, "Electricity and Magnetism," McGraw-Hill Inc., New York, N. Y., p. 38; 1938.



(a) TWO SINGLE-WIRE TRANSMISSION LINES USING A COMMON GND. PLANE



(b) TWO SINGLE-WIRE TRANSMISSION LINES WITH IMAGES

Fig. 7—Two single-wire transmission lines with images.

Fig. 8 is a photograph of an experimental set up for such a two-wire system. On Fig. 9, theoretical curves and experimental points representing Coupling vs Frequency with line length as a parameter are shown. The experimental points verify the theory.

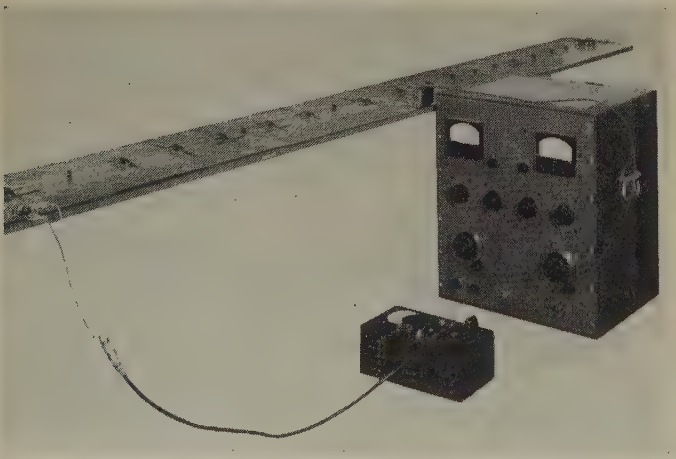


Fig. 8

B. A Three-Wire System

The characteristic equation for a four-wire system under light-coupling conditions is given by (1). It is consistent with field theory to state that M and C_m under all coupling conditions under consideration involve the distances D_{14}, D_{23}, D_{13}, and D_{24} in the same way; hence the characteristic equation is independent of these distances. A three-wire directional coupler is a special case of a four-wire coupler for which either D_{13} or D_{24} approach zero. We would expect, therefore, a three-wire system as shown in Fig. 10 to exhibit directional coupling characteristics.

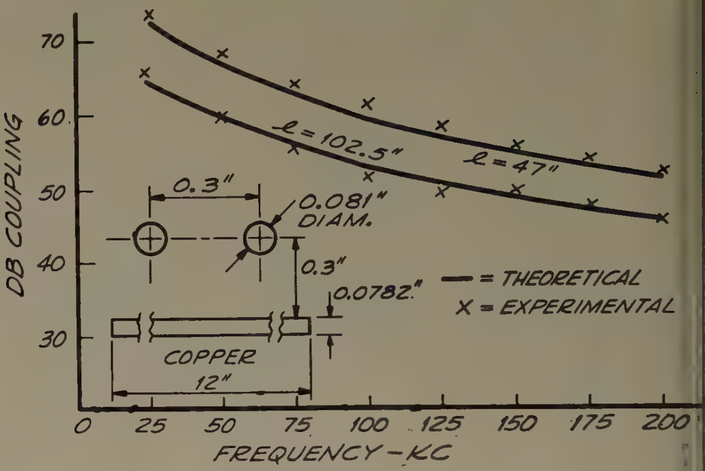


Fig. 9—Db coupling vs frequency.

There are two distinct methods of solving for the coupling of this three-wire system. The *first method* consists of finding the size and position of the images of the active wires in the grounded common wire.

These images may be found readily by the use of the conformal transformation determined by

$$W = \frac{1}{Z}. \tag{27}$$

This equation transforms circles into straight lines or circles. It is still, of course, necessary to leave the original configuration unaltered and merely replace the common wire with the proper images. Furthermore, the radius of the common wire must be multiplied by a proper factor so that it becomes unity, before the transformation is applied.

Reference is made to Fig. 10. Variation of this method is to transform the common wire into the upper half-plane by the conformal transformation:

$$W = i\frac{1-Z}{1+Z}, \tag{28}$$

and, by using the inverse transformation

$$Z = \frac{i-W}{i+W}, \tag{29}$$

to transform back to the original configuration, except that the common wire has been replaced by two images whose size and position are completely known. This approach requires a knowledge of the impedance of unsymmetrical two-wire lines which is given by Schelkunoff.[16]

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[16] S. A. Schelkunoff, "Electromagnetic Waves," D. Van Nostrand Book Co., Inc., New York, N. Y., p. 179; 1943.

It is quite apparent from Fig. 10 that the new D_{24} dimension obtained by any of the image methods is very small; hence we would expect a much higher degree of coupling for such a three-wire system than was found for an open four-wire system of comparable dimensions. This is, in fact, the case. For the arrangements attempted, between 10 and 20 db more coupling was experienced.

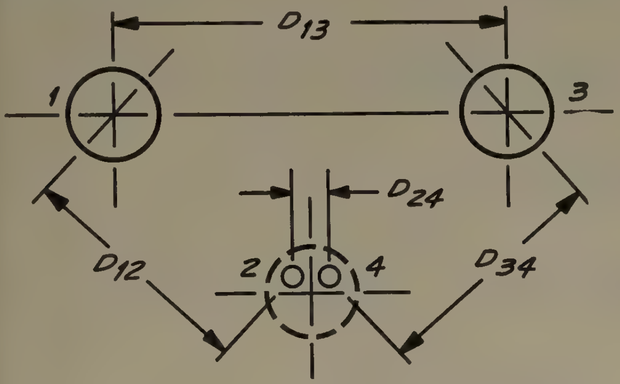

Fig. 10—3-wire directional coupler with images shown.

The *second method* involves conformally mapping the three-wire system into the configuration of Fig. 11 by (28). Once this is done, the theory of images is applied and a four-wire system is arrived at. In this method of attack we do not transform back to a system as shown in Fig. 10 but apply our four-wire directional coupler equations directly.

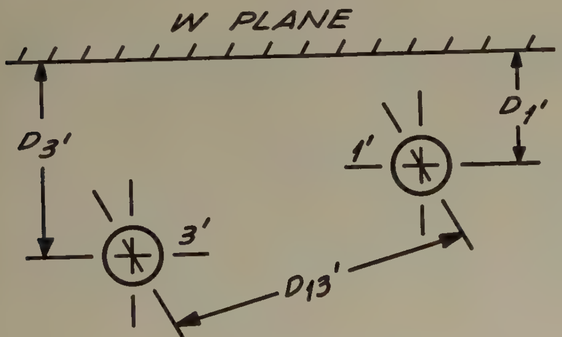

Fig. 11—Transformed 3-wire directional coupler.

It has been shown by Spangenberg[17] that, for a TEM mode on an open two-wire transmission line, LaPlace's Equation is satisfied, and that conformal mappings may be made with complete accuracy; consequently the above transformation is valid.

Fig. 12 corroborates the preceding theory and provides for a comparison of a three-wire directional coupler with a comparable four-wire system. A 15.5 db improvement in coupling is observed for the three-wire system over the four-wire system. The calculation of the theoretical curves by any of the previously discussed methods yields identical results. As further verification of the preceding theory, another configuration was tested wherein the center wire of a three-wire system was grounded. The experimental points were in even closer agreement with theory than for Fig. 12.

It can readily be seen that, whenever it is possible to use a three-wire system instead of a four-wire system, the three-wire system should be utilized because of the tremendous increase in coupling which results for a given frequency and coupling length.

In order to utilize a three-wire system for power lines and not have the radio frequency carrier equipment at the high potential of the power line, it becomes necessary to use a high voltage capacitor. This capacitance grounds the 60-cycle power line for radio frequencies only, and thereby permits the carrier transmitter and associated equipment to be operated at ground potential.

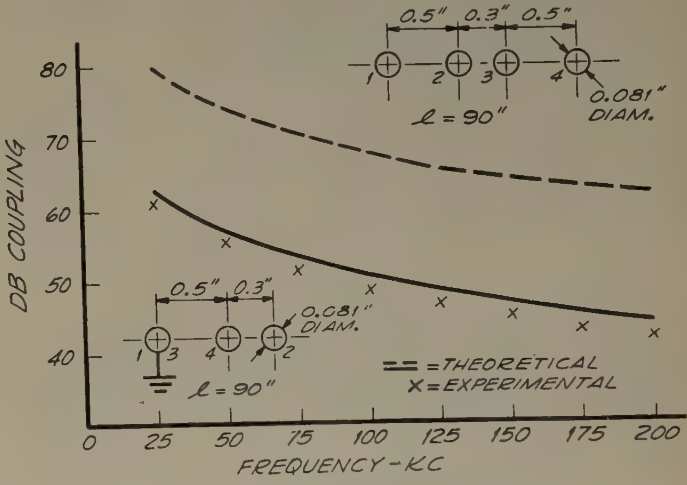

Fig. 12—Db coupling vs frequency.

C. A Four-Wire System with Two Wires Grounded

Let us consider the symmetrical case shown in Fig. 13, in which the two outer wires are grounded. By the method of images, we may replace one of the ground wires by the images of the two active wires. Any of the methods of the preceding section may be used to find the images.

In order to replace the ground wire by its images we must insist that the images carry $-i_4$ and $-i_1$, respectively. The sum of the return currents is now i_1+i_4, and hence all of the current is accounted for; consequently the effect of the second ground wire may be considered as nil. It is clear, however, that this system is not identical to the three-wire system insofar as Z_{01} is now the impedance between wires (1) and (2), whereas if wire (2) were removed Z_{01} would be the impedance between wires (1) and (3). The problem, therefore, reduces to one of finding the size and position of the images in one of the ground wires.

It can readily be seen that our problem has been reduced to that of a Three-Wire Directional Coupler, which has been completely solved and verified in the preceding section. An experimental check on the symmetrical four-wire system is shown in Fig. 13 and the amount of coupling is obviously in excellent agreement with the theory. For the unsymmetrical case a little more consideration would be required.

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[17] K. Spangenberg, "The use of conformal transformations in ultra high frequency transmission line problems," Proc. I.R.E., vol. 37, p. 532; May, 1949.

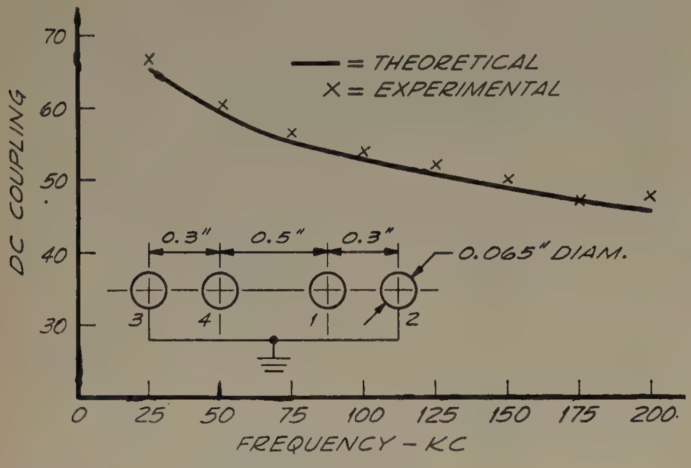

Fig. 13—Db coupling vs frequency.

D. A Shielded Two-Wire Directional Coupler

A fifth configuration as shown in Fig. 14 was analyzed and found to exhibit directional properties. Wire (1) and the grounded shield represent line (1, 2), whereas wire (3) and the grounded shield represent line (3, 4).

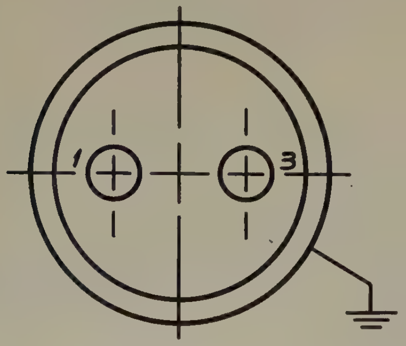

Fig. 14—Shielded two-wire directional coupler.

As before we may use the method of images by imaging the wires (1) and (3) into the shield. The problem, therefore, reduces to one of finding the size and position of the images outside the grounded shield. The same mathematics may be applied in either case, except that, instead of imaging two active wires into a ground wire, we have to image two active wires onto the outside of a ground wire.

If the ground shield is replaced by the images of the two active wires, Fig. 15 results. Fig. 16 is presented as verification of the above theory. In Fig. 16 the images are on opposite sides of the vertical center-line, whereas in another configuration which was verified the images are on the same side of the vertical center-line. Experimental directivities with this system can be made to approach infinity over the entire tested frequency range. Directivities in excess of 50 db are not too difficult to achieve experimentally.

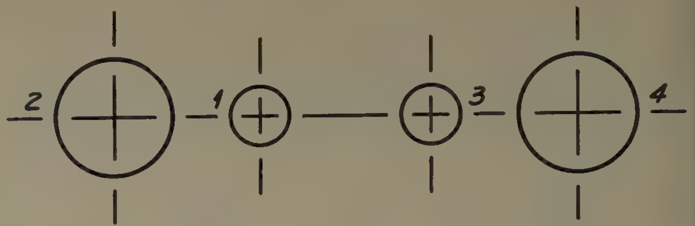

Fig. 15—Transformed shielded two-wire directional coupler.

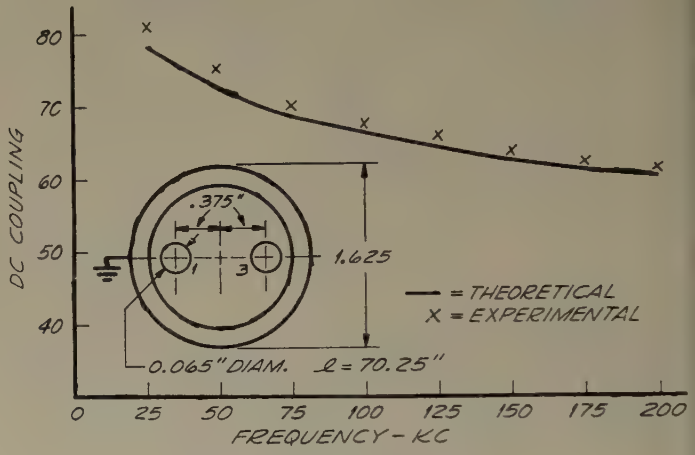

Fig. 16—Db coupling vs frequency.

LUMPED CIRCUIT DIRECTIONAL COUPLERS

If one considers that the coupling be done by lumped parameters, i.e., M and C_m are lumped values, it is possible to arrive at the configuration of Fig. 17. This coupler will conform to the basic requirements of the characteristic equation whether one, two, or no transmission lines are involved. Directivities greater than 45 db and couplings of less than -1 db are easily achievable with it. Fig. 18 shows a typical input impedance plot versus frequency for such a coupler.

CONCLUSIONS

Analytical expressions have been derived for the directivity and coupling with and without reflections being present for six different types of transmission line systems. One of the most significant results has been the discovery which shows that, if certain impedance relations are maintained, it is possible to mismatch completely the transmission lines and still obtain infinite directional coupling.

The analysis of directional coupling with reflections present on both lines has permitted us to write a more general scattering matrix than has been done heretofore.

Each of the directional couplers developed in this paper has certain advantages, namely:

1. The four-wire system is most desirable for high-voltage systems where voltage breakdown is a factor.

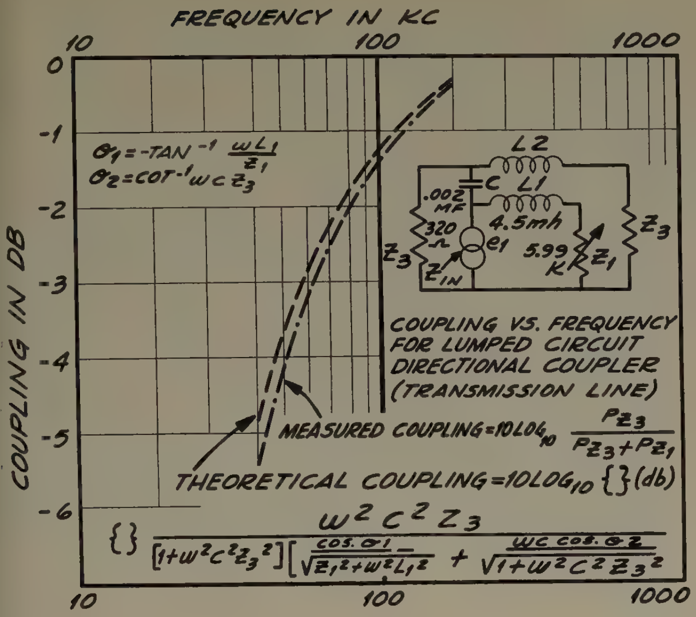

Fig. 17—Lumped circuit coupler.

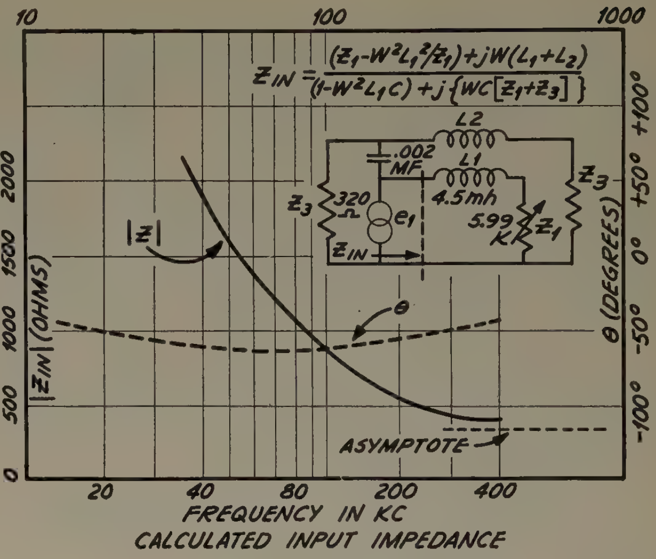

Fig. 18

2. The two-wire system requires the least wire and is therefore desirable where the use of a ground plane is permissible.
3. The four-wire system with two wires rounded is most desirable where the wires must be high off the ground, and the impedance of each line is desired to be low. This arrangement yields much higher coupling than either of the above systems but requires two of the wires to be at a common potential.
4. The three-wire system is best suited for applications where one wire may be used as a common wire and the coupling is desired to be high.
5. The shielded two-wire system has practically no noise or other extraneous voltage and hence is desirable whenever noise or pick-up voltage is important.
6. The lumped circuit coupler has the highest amount of coupling, requires practically no coupling length and does so with a high directivity.

One of the most desirable features of the open-wire transmission line direction couplers is its relative insensitivity to variations in frequency (i.e.: its broad bandedness). If coupling sections of one quarter wavelength at the center frequency are utilized, then a frequency range of approximately three to one will yield only 3 db of coupling variation.

It is to be expected that other wire transmission line configurations will satisfy the "characteristic equation" and hence display directional characteristics. The extension of the art of directional coupling to the lower radio frequencies should prove of great value.

Although most of the tests have been performed at relatively low radio frequencies, several of the couplers were tested at 30 mc. and operated in an entirely satisfactory way. Also consideration of the basic equations for M and C_m shows that relative dielectric constants other than 1 make no difference as regards directivity, and hence dielectric-surrounded wire directional couplers are feasible and practical.

APPENDIX I

Derivation of New Scattering Matrix

In order to conform with standard directional coupler definitions, it is necessary to describe directional couplers as having no coupling between terminals (1) and (3), and (2) and (4); therefore this will be done for the following section on matrixes only.

Because of the preceding developments on mismatched transmission line directional couplers and the subsequent additional knowledge, it is possible to formulate a more general directional coupler definition and a more general scattering matrix than has been possible in the past.

The scattering matrix of a four-terminal network is usually given by

$$S = \begin{pmatrix} S_{11} & S_{12} \cdots S_{1n} \\ S_{21} & S_{22} \cdots S_{2n} \\ \cdot & \\ \cdot & \\ \cdot & \\ S_{n1} & \cdot \quad \cdots S_{nn} \end{pmatrix} \tag{30}$$

where the scattering coefficient, S_{mn}, is a measure of the amplitude of the wave scattered into the mth guide by an incident wave of unit amplitude in the nth guide. In particular, therefore, the coefficient S_{mn} represents the reflection coefficient at the terminal T_m when all other terminals are matched. It is well known that the scattering matrix is symmetrical and unitary for non-dissipative structures, that is,

$$S_{mn} = S_{nm} \tag{31}$$

and

$$\sum_{\beta=1}^{n} S_{\beta M}{}^{*} S_{\beta n} = \begin{cases} 1 & \text{if} \quad m = n \\ 0 & \text{if} \quad m \neq n. \end{cases} \tag{32}$$

If the given structure possesses geometrical symmetries, it is possible to derive corresponding symmetry relations among the scattering coefficients.

If a junction is "completely matched," the diagonal elements $S_{mm}=0$.

The relationship between the scattering matrix above and the impedances of the network may be shown by used of the following relationship:

$$S = (Z - I)(Z + I)^{-1}, \tag{33}$$

where Z is the impedance matrix of the network and I is the unit matrix. The term $(Z+I)^{-1}$ is the inverse of the matrix $(Z+I)$.

If a four-terminal-pair network has certain physical characteristics and if each junction is terminated in a prescribed manner, it may display directional characteristics. These directional properties may best be seen by considering the directional coupler definition given by the author of this paper.

The scattering matrix for a four-terminal-pair network with no coupling between junctions (1) and (3) and between (2) and (4) is given by

$$S = \begin{pmatrix} S_{11} & S_{12} & 0 & S_{14} \\ S_{21} & S_{22} & S_{23} & 0 \\ 0 & S_{32} & S_{33} & S_{34} \\ S_{41} & 0 & S_{43} & S_{44} \end{pmatrix}. \tag{34}$$

The above matrix will represent a directional coupler, if the following conditions are met:

(a) The region of coupling is small compared to a wavelength,

(b) $Z_{44}Z_{22}=1$, (35)

and

(c) $Z_{33}Z_{11}=1$. These conditions have been derived in the text.

Eq. (32) may be put into a more customary form, if the conditions that the scattering matrix be symmetrical and unitary are utilized. From the unitary conditions, (32), the following relationships may be written:

$$\begin{aligned} &\text{(a)} \quad |S_{11}|^2 + |S_{12}|^2 + |S_{14}|^2 = 1, \\ &\quad \vdots \\ &\text{(d)} \quad |S_{41}|^2 + |S_{43}|^2 + |S_{44}|^2 = 1, \end{aligned} \tag{36}$$

and

$$\begin{aligned} &\text{(a)} \quad S_{11}^*S_{12} + S_{21}^*S_{22} = 0, \\ &\quad \vdots \\ &\text{(e)} \quad S_{12}^*S_{11} + S_{22}^*S_{21} = 0. \end{aligned} \tag{37}$$

All of the above equations are not independent.

Solving the above equations for S_{44}^* and S_{11}^*, respectively, we obtain

$$\begin{aligned} &\text{(a)} \quad S_{44}^* = - S_{14}^*S_{11}/S_{14}, \\ &\text{(b)} \quad S_{11}^* = - S_{14}^*S_{44}/S_{14}. \end{aligned} \tag{38}$$

There is a great deal of arbitrariness in the phases of the elements S. This indeterminateness can be eliminated by the proper choice of reference planes. As an example, the location of the terminal plane (4) may be chosen in such a way that the S_{14} is positive imaginary. If this is done, we may write the solutions of (38) directly,

$$S_{11} = \alpha_1 + j\beta_1, \qquad S_{44} = \alpha_1 - j\beta_1, \tag{39}$$

provided

$$S_{14} = j\beta_2. \tag{40}$$

Similarly with the aid of the other equations, the following relationships may be obtained: $S_{11}=\alpha_1+j\beta_1$, $S_{22}=\alpha_1-j\beta_1$, if

$$S_{12} = j\beta_3,$$

and

$$S_{22} = \alpha_1 - j\beta_1, \qquad S_{33} = \alpha_1 + j\beta_1,$$

provided

$$S_{23} = j\beta_4. \tag{41}$$

Furthermore, it can readily be shown that

$$S_{11}=S_{33}, S_{44}=S_{22}, |S_{23}| = |S_{14}|, |S_{12}| = |S_{34}|, \beta_3=\beta_4,$$

and

$$S_{34} = - j\beta_4. \tag{42}$$

For conciseness we let

$$\bar{\gamma} = \alpha_1 + j\beta, \quad \text{and} \quad \bar{\gamma}^* = \alpha_1 - j\beta_1. \tag{43}$$

With the choice of reference planes as indicated above, the scattering matrix for our directional coupler becomes

$$S = \left(\begin{array}{cc:cc} \bar{\gamma} & j\beta_3 & 0 & j\beta_2 \\ j\beta_3 & \bar{\gamma}^* & j\beta_2 & 0 \\ \hdashline 0 & j\beta_2 & \bar{\gamma} & -j\beta_3 \\ j\beta_2 & 0 & -j\beta_3 & \bar{\gamma}^* \end{array}\right). \tag{44}$$

In the customary definition of a directional coupler, it has been necessary to specify that every directional coupler is a "completely matched" unit with all lines terminated in their characteristic impedances. The condition that a directional coupler be "completely matched" is a *sufficient*, but not a *necessary* condition insofar as mismatched sections may be directional couplers also, as evidenced by the development with respect to the author's definition.

ACKNOWLEDGMENT

Credit is due to H. Magnuski and J. F. Byrne of Motorola Inc., who made this investigation possible, and to Dr. R. E. Beam of Northwestern University, who offered many helpful criticisms and suggestions.

The Power Spectrum of a Carrier Frequency Modulated by Gaussian Noise*

JOHN L. STEWART†, MEMBER, IRE

Summary—**General expressions for the power spectra of carriers frequency and phase modulated by Gaussian noise are given as functions of the shapes of the power spectra of the modulating voltages. Specific asymptotic (closed-form) expressions are obtained for the special case when the modulating voltage has a rectangular low-pass power spectrum. This particular case corresponds to the use of maximally flat and similar video amplifier tuning schemes. The half-power bandwidths of the power spectra of the modulated carriers are obtained from these asymptotic expressions and are plotted as functions of the root-mean-square frequency and phase deviations. The delta function at the carrier frequency (representing power remaining in the carrier) is evaluated for phase and frequency modulation where, in the latter case, the delta function exists only if the power spectrum of the modulating voltage does not extend to zero frequency.**

INTRODUCTION

A KNOWLEDGE of the power spectra of phase- and frequency-modulated signals is of considerable interest in the design of networks intended to accommodate these signals—in particular, when the modulating voltage consists of random Gaussian noise. In addition to interest in such signals *per se*, Gaussian noise is often a reasonable "signal" to assume as a replacement for many types of complex modulating waves such as those of voice and television.

Many types of oscillators are subject to both amplitude and frequency fluctuation noise—in some cases, frequency modulation noise may have far greater significance as, for example, in magnetrons[1] and some reactance modulated oscillators. Frequency modulation noise in such devices is similar to that obtained by frequency modulating an oscillator with a low-pass band of Gaussian noise where the root-mean-square frequency deviation is much less than the modulator bandwidth. The result is a continuous power spectrum having skirts that fall off only as the square of the frequency difference from the carrier. If applicable to the local oscillator of a superheterodyne receiver, the effect of such a noise spectrum is obvious.

It should not be inferred that a knowledge of the power spectrum of a frequency- or phase-modulated carrier tells a great deal. In fact, the power spectrum of FM or ϕM yields much less (relative) data than that of an amplitude-modulated wave. For example, if the frequency deviation of an FM signal is larger than the bandwidth of the modulating voltage, the shape of the spectrum is essentially independent of the bandwidth of the modulating signal. Yet a knowledge of whether the bandwidth of the modulating signal is, say, one megacycle per second rather than one cycle per year, is of obvious importance.

Nevertheless, an important (albeit incomplete) piece of information relating to frequency- or phase-modulated signals (modulated by Gaussian noise) is the power spectrum. It is the spectra of such signals that are of concern here. The problem solved in this report was solved in part by Middleton.[2-4] The approach given here is a modification of Middleton's with the resulting answers applying to somewhat different situations. Middleton did not treat the case of a rectangular modulating signal power spectrum—he solved for the Gaussian case. Actually, the rectangular spectrum more closely corresponds to the use of maximally flat, shunt-peaked, and similar amplifier tuning schemes. Middleton presented his results in the form of power spectrum curves computed from infinite series. Here asymptotic (closed form) expressions are obtained with the half-power bandwidth of the power spectrum ultimately appearing graphically.

BASIC EQUATIONS

It is the purpose of this section to set down some of the fundamental relations originally obtained by Middleton and at a later date by the author using a slightly different mathematical approach.[5]

The voltage of an FM or ϕM wave is given by

$$V(t) = A_0 \exp \{ j[\omega_0 t + \psi(t)] \} \tag{1}$$

where A_0 is the peak amplitude, ω_0 is the (radian) carrier frequency, and $\psi(t)$ is the instantaneous phase-angle shift created by the modulation. In FM,

$$\psi(t) = \int^{t} D(t')dt', \tag{2}$$

and in ϕM

$$\psi(t) = D(t), \tag{3}$$

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† University of Michigan, Ann Arbor, Mich.

[1] J. L. Stewart, "Theory of Frequency Modulation Noise in Tubes Employing Phase Focusing," Technical Report No. 28, Electronic Defense Group, University of Michigan, Ann Arbor, Mich.; March, 1954.

[2] D. Middleton, "On the Distribution of Energy in Noise- and Signal-Modulated Waves," Technical Report No. 99, Cruft Laboratory, Harvard University, Cambridge, Mass.; March, 1950.

[3] D. Middleton, "The distribution of energy in randomly modulated waves," *Phil. Mag.*, Ser. 7, p. 689; July, 1951.

[4] D. Middleton, "On the distribution of energy in noise and signal modulated waves," *Quart. Appl. Math.*, Part I, vol. 9; January, 1952, and Part II, vol. 10; April, 1952.

[5] J. L. Stewart, "The Power Spectrum of a Carrier Phase or Frequency Modulated by Gaussian Noise," Technical Report No. 23, Electronic Defense Group, University of Michigan, Ann Arbor, Mich.; 1953.

where $D(t)$ is the modulating voltage. If a normalized modulating voltage $V_n(t)$ is defined such that

$$\overline{[V_n(t)]^2} = 1, \qquad \overline{[V_n(t)]} = 0, \tag{4}$$

the total modulating voltage becomes

$$D(t) = DV_n(t) \tag{5}$$

where D^2 is the mean-square deviation (from the carrier frequency) of the FM or the mean-square phase shift for ϕM. It is also the mean-square value of the modulating voltage, if frequency and voltage are linearly related.

The autocorrelation function of the modulated carrier $R(\tau)$ can be obtained using a straightforward mathematical attack as done by Middleton to obtain

$$R(\tau) = (A_0^2/2) \cos \omega_0\tau \cdot \exp \left\{ - \left[\overline{\psi(t)^2} - \overline{\psi(t)\psi(t+\tau)} \right] \right\} \tag{6}$$

which applies to either FM or ϕM.

With some difficulty, $\overline{\psi(t)}$ can be expressed in terms of the autocorrelation function of the modulating voltage $R_m(\tau)$. Then, the Wiener-Khintchine theorem can be applied to find the normalized power spectrum for frequency modulation $W_F(\Delta\omega)$ as

$$W_F(\Delta\omega) = \frac{A_0^2}{2\pi} \int_0^\infty \cos \Delta\omega\tau \cdot \exp \left\{ - D^2 \int_0^\infty \frac{W_m(\omega)}{\omega^2} (1 - \cos \omega\tau) d\omega \right\} d\tau \tag{7}$$

where $W_m(\omega)$ is the normalized power spectrum of the modulating voltage $V_n(t)$ and where $\Delta\omega$ is the difference frequency from the unmodulated carrier. The corresponding power spectrum for phase modulation $W_\phi(\Delta\omega)$ is the same as (7) except that the factor $1/\omega^2$ under the integral in the exponent does not appear.

The only essential difference between FM and ϕM is the factor $1/\omega^2$. By taking an FM wave modulated with a voltage having the power spectrum $W_m(\omega)$ and changing the spectrum by means of a linear filter to $\omega^2 W_m(\omega)$ (a differentiating network), phase modulation is obtained. The converse also applies. The close relation between FM and ϕM is notable.

FM SPECTRUM WITH RECTANGULAR MODULATION POWER SPECTRUM

At this point, a specific (normalized) power spectrum can be assumed for $W_m(\omega)$ in (7) which leads to a specific expression for the power spectrum of a frequency modulated wave. A rectangular power spectrum will be assumed for the modulating voltage. This choice will simplify some of the relations while, at the same time, it is a fair approximation to many types of wide-band amplifier functions.

Let $W_m(\omega)$ be given by

$$W_m(\omega) = \begin{cases} \dfrac{1}{B}, & 0 < \omega < B \\ 0, & \omega > B \end{cases} \tag{8}$$

where B is the radian bandwidth of the modulating-voltage power spectrum. Using (8) in (7) results in

$$W_F(\Delta\omega) = \frac{A_0^2}{2\pi} \int_0^\infty \cos \Delta\omega\tau \cdot \exp \left\{ - \frac{D^2\tau}{B} \int_0^{B\tau/2} \left[\frac{\sin y}{y} \right]^2 dy \right\} d\tau. \tag{9}$$

Changing the variable $B\tau$ to x, expanding $[(\sin y)/y]^2$, and integrating termwise,

$$W_F(\Delta\omega) = \frac{A_0^2}{2\pi B} \int_0^\infty \cos \frac{\Delta\omega x}{B} \cdot \exp \left\{ - \frac{D^2}{B^2} \left[\frac{x^2}{1.2!} - \frac{x^4}{3.4!} + \cdots \right] \right\} dx. \tag{10}$$

For large D/B, the exponential is appreciable only for small x; thus, the asymptotic case exists:

$$W_F(\Delta\omega) \cong \frac{A_0^2}{2\pi B} \int_0^\infty \cos \frac{\Delta\omega x}{B} \exp \left[- \frac{D^2 x^2}{2B^2} \right] dx. \tag{11}$$

This integral is well known with the result

$$W_F(\Delta\omega) \cong (A_0^2/2) \frac{\exp(-\Delta\omega^2/2D^2)}{(2\pi D^2)^{1/2}} \tag{12}$$

which is the Gaussian function. It is to be noted that the carrier has been smeared entirely into a continuum and the modulator bandwidth does not appear as a parameter.

The Gaussian nature of the power spectrum is a more general consequence of large D/B than might at first be thought. In fact, by inspecting the nature of the expansion of (10), almost any "ordinary" sort of low-pass spectrum shape for $W_m(\omega)$ leads to the Gaussian power spectrum for $W_F(\Delta\omega)$.

The relative half-power point of (12) is given when $W_F(\Delta\omega)$ is half that at $\Delta\omega = 0$. If the frequency yielding this is called B_F, there is obtained

$$B_F = D\sqrt{2 \ln 2} = 1.18D \tag{13}$$

which is one of the asymptotic expressions plotted in Fig. 1.

At the other extreme is the spectrum $W_F'(\Delta\omega)$ as $D/B \rightarrow 0$. In this case, inspection of (9) shows that only a small part of the total value of the integral in the exponent is made up of parts for which $B\tau/2$ is small. In the limit, the only significant contributions occur for large $B\tau/2$. Thus, the integral of $[(\sin y)/y]^2$ can be approximated by its asymptotic value of $\pi/2$ to give

$$W_F'(\Delta\omega) \cong \frac{A_0^2}{2\pi B} \int_0^\infty \cos \frac{\Delta\omega x}{B} \exp \left[- \frac{\pi D^2 x}{2B^2} \right] dx. \tag{14}$$

This integral is also known with the result

$$W_F'(\Delta\omega) \cong (A_0^2/2) \frac{D^2/2B}{(\pi D^2/2B)^2 + (\Delta\omega)^2} \tag{15}$$

which is the same as the output of a simple resistance-capacitance filter with white noise at the input.

The half power point of (15) occurs for a $\Delta\omega = B_F'$ of

$$B_F' = \pi D^2/2B = 1.57 D^2/B. \tag{16}$$

Eq. (16) is also plotted in Fig. 1. In this figure, the dotted line shows the "estimated" true curve.

Eqs. (13) and (16) yield the same value at the "crossover" point for a D/B of

$$[D/B]_c = (2/\pi)\sqrt{2 \ln 2} = 0.751. \tag{17}$$

As a matter of classification, the usual frequency-modulated signal can be said to persist for $D/B > 0.751$, whereas, for $D/B < 0.751$, the condition approaches that of narrow-band frequency modulation.

An interesting effect also noted by Middleton occurs when the power spectrum of the modulating signal is zero at very low frequencies—if the spectrum is constant for $B_1 < \Delta\omega < B$ and zero otherwise, it has the form of a rectangular band-pass rather than low-pass spectrum. If B/B_1 is large, the continuous part of the power spectrum of the modulated carrier will be little changed; however, a small delta function will be introduced at the carrier frequency having the form

$$(A_0^2/2) \exp(-2D^2/B_1 B)\delta(\Delta\omega).$$

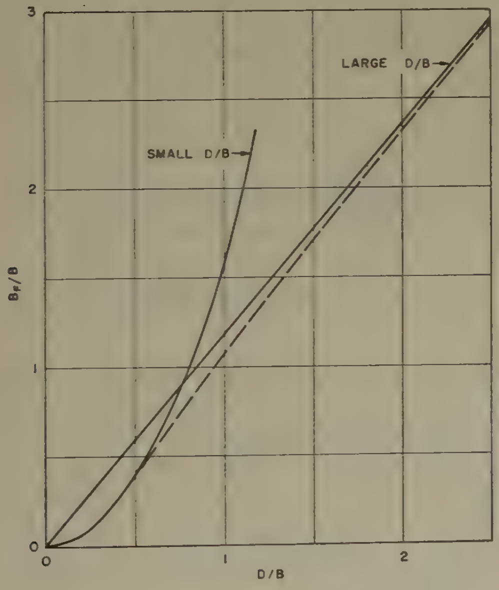

Fig. 1—Half bandwidth with frequency modulation.

ϕM Spectrum with Rectangular Modulation Power Spectrum

In this section, the power spectrum of a phase-modulated carrier with a rectangular modulation power spectrum will be obtained in close analogy to that done for the case of the frequency-modulated carrier. Substituting the modulation spectrum as given by (8) into (7) with the factor $1/\omega^2$ in (7) deleted, one obtains

$$W_\phi(\Delta\omega) = \frac{A_0^2}{2\pi}\int_0^\infty \cos \Delta\omega\tau \cdot \exp\left[-\frac{4D^2}{B\tau}\int_0^{B\tau/2} \sin^2 y\,dy\right] d\tau. \tag{18}$$

Using a change of variable, this expression becomes

$$W_\phi(\Delta\omega) = \frac{A_0^2}{2\pi B}\int_0^\infty \cos\frac{\Delta\omega x}{B} \cdot \exp\left[-\frac{4D^2}{x}\int_0^{x/2} \sin^2 y\,dy\right] dx. \tag{19}$$

The integral in the exponent can be evaluated to give

$$W_\phi(\Delta\omega) = \frac{A_0^2}{2\pi B}\int_0^\infty \cos\frac{\Delta\omega x}{B} \cdot \exp\left[-D^2\left(1 - \frac{\sin x}{x}\right)\right] dx. \tag{20}$$

At $\Delta\omega = 0$, the integrand of (20) does not converge. This indicates the presence of a delta function at $\Delta\omega = 0$. It is necessary to remove this singularity from the integral before the continuous part of the spectrum may be obtained.

The number $\exp(-D^2)$ may be added and subtracted in a factor containing the exponential in (20) as

$$W_\phi(\Delta\omega) = \frac{A_0^2}{2\pi B}\int_0^\infty \cos\frac{\Delta\omega x}{B}\left\{\exp(-D^2) - \exp(-D^2) + \exp\left[-D^2\left(1 - \frac{\sin x}{x}\right)\right]\right\} dx \tag{21}$$

which gives

$$\begin{aligned} W_\phi(\Delta\omega) = {} & \frac{A_0^2}{2\pi B}\exp(-D^2)\int_0^\infty \cos\frac{\Delta\omega x}{B}\,dx \\ & + \frac{A_0^2}{2\pi B}\int_0^\infty \cos\frac{\Delta\omega x}{B} \cdot \left\{\exp\left[-D^2\left(1 - \frac{\sin x}{x}\right)\right] - \exp(-D^2)\right\} dx. \end{aligned} \tag{22}$$

The first of these integrals is related to the half delta function and the second is convergent at $\Delta\omega = 0$ which indicated that it gives the continuous part of the power spectrum. Thus,

$$\begin{aligned} W_\phi(\Delta\omega) = {} & \frac{A_0^2}{2}\exp(-D^2)\delta(\Delta\omega) \\ & + \frac{A_0^2}{2\pi B}\int_0^\infty \cos\frac{\Delta\omega x}{B} \cdot \left\{\exp\left[-D^2\left(1 - \frac{\sin x}{x}\right)\right] - \exp(-D^2)\right\} dx. \end{aligned} \tag{23}$$

For large D, the power remaining in the carrier becomes quite small and can often be neglected.

An interesting observation concerning FM and ϕM (also noted by Middleton), can be made by comparing (23) and (15). As the magnitude of the modulating voltage in FM is decreased, the spectrum becomes narrower and narrower without limit, finally resulting in a delta function. However, the spectrum is continuous at all times. In phase modulation, on the other hand, reducing the magnitude of the modulating voltage removes power from the side bands and places it in the ever present delta function at the carrier frequency, causing the delta function to increase in included area.

The second term of (23) gives the continuous part of the spectrum. For large D, the integrand of (23) is appreciable only for small x. In this case

$$\begin{aligned}\exp\left[-D^2\left(1-\frac{\sin x}{x}\right)\right]&-\exp(-D^2)\\ &=\exp\left[-D^2\left(\frac{x^2}{3!}-\frac{x^5}{5!}+\cdots\right)\right]\\ &\quad-\exp(-D^2)\cong\exp\left(-\frac{D^2x^2}{6}\right). \qquad (24)\end{aligned}$$

Using this in (23) and performing the integration,

$$W_\phi(\Delta\omega)\cong\frac{A_0{}^2}{2}\left[\exp(-D_2)\delta(\Delta\omega)+\frac{\exp\left(-\dfrac{\Delta\omega^2}{2D^2B^2/3}\right)}{(2\pi D^2B^2/3)^{1/2}}\right]. \qquad (25)$$

The half power point of the (again Gaussian) continuous part of the power spectrum occurs at a $\Delta\omega = B_\phi$ of

$$B_\phi = DB\sqrt{\frac{2\ln 2}{3}} = 0.68DB. \qquad (26)$$

Eq. 26 is one of the asymptotic expressions plotted in Fig. 2.

For small D, there exist important contributions for large values of x in (21) as well as for small values. However, $D^2(\sin x)/x$ is always small; hence, the series expansion for the exponential may be used to advantage to give

$$\begin{aligned}W_\phi(\Delta\omega)&=\frac{A_0{}^2}{2}\exp(-D^2)\delta(\Delta\omega)\\ &\quad+\frac{A_0{}^2D^2}{2\pi B}\exp(-D^2)\int_0^\infty\cos\frac{\Delta\omega x}{B}\\ &\quad\cdot\left[\frac{\sin x}{x}+\frac{D^2}{2!}\left(\frac{\sin x}{x}\right)^2\right.\\ &\quad\left.+\frac{D^4}{3!}\left(\frac{\sin x}{x}\right)^3+\cdots\right]dx. \qquad (27)\end{aligned}$$

The integral of (27) becomes a sum of integrals. The first order term is a known integral,

$$\int_0^\infty\frac{\sin x\cos\dfrac{\Delta\omega x}{B}}{x}\,dx=\begin{cases}0, & \left(\dfrac{\Delta\omega}{B}\right)^2>1\\ \dfrac{\pi}{4}, & \left(\dfrac{\Delta\omega}{B}\right)^2=1\\ \dfrac{\pi}{2}, & \left(\dfrac{\Delta\omega}{B}\right)^2<1.\end{cases} \qquad (28)$$

To a first order, the spectrum consists of a delta function at $\Delta\omega=0$ and a continuous and flat spectrum over the range of frequencies occupied by the modulating signal. For this asymptotic case, bandwidth of continuous part of the spectrum is constant at $B_\phi'=B$ and plotted as one of the asymptotic expressions in Fig. 2.

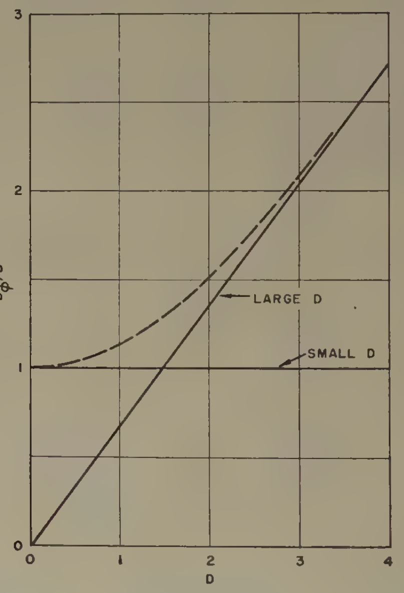

Fig. 2—Half bandwidth with phase modulation.

The spectrum for small D can thus be written

$$W_\phi(\Delta\omega)\cong\begin{cases}\dfrac{A_0{}^2}{2}\left[\exp(-D^2)\delta(\Delta\omega)\right.\\ \qquad\left.+\dfrac{D^2}{2B}\exp(-D^2)\right], & \Delta\omega<B.\\ 0, & \Delta\omega>B\end{cases} \qquad (29)$$

A second-order approximation for the continuous part of the spectrum for small D can be obtained by including the second integral of the series of (27). Only the general effect will be noted here. The function $(\sin x/x)^2$ can be approximated very roughly with the function $a/(1+x^2)$, in which case, the correction term becomes an exponential function which is to be added to the rectangular first order spectrum. The approximate correction term is $(A_0{}^2/2)\,(aD^2/2B)\exp(-\Delta\omega/B)$.

Step Discontinuities in Disk Transmission Lines*

R. N. BRACEWELL†

Summary—**Step discontinuities in disk transmission lines may be approximately allowed for using the known results for steps in parallel plane transmission lines. A correction factor is calculated for the case of a disk line and shown to depend on whether the step up is away from or towards the center. The factor is given by**

$$\frac{\displaystyle\sum_1^\infty \frac{K_1(n\pi\beta)}{K_0(n\pi\beta)}\,\frac{\sin^2 n\pi\alpha}{\alpha^2 n^3}}{\displaystyle\sum_1^\infty \frac{\sin^2 n\pi\alpha}{\alpha^2 n^3}}$$

for an outward step up, where K_0 and K_1 are modified Bessel functions of the second kind. For an inward step up, K_0 and K_1 are replaced by I_0 and I_1, the modified Bessel functions of the first kind. Graphical charts are given which allow the factors to be directly determined from the shape parameters α and β.

INTRODUCTION

A SIMPLE DISK transmission line (or radial transmission line) is shown in cross-section in Fig. 1. The flow of current in the two conductors is along the radii and the lines of electric force are normal to the two conductors, the excitation and loading of the line being supposed circularly symmetrical. Disk transmission lines gain their importance from the frequency with which they occur in microwave apparatus. Formulas and design charts for handling them have been given by Marcuvitz.[1]

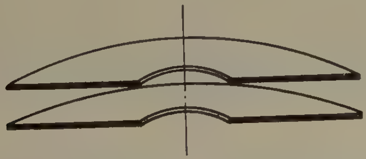

Fig. 1—Simple disk transmission line.

Figs. 2 and 3 show tandem arrangements of two disk transmission lines with different spacings. We shall call arrangements of this kind stepped disk lines. This paper is concerned with calculating the effect of such steps and with presenting the results in a form that makes them easy to utilize. A considerable variety of apparatus incorporates disk lines with steps, a few examples being shown in Fig. 4.

The mathematical procedure for dealing with steps in parallel plane transmission lines has been given by Whinnery and Jamieson,[2] and has been extended by Whinnery, Jamieson and Robbins[3] to the case of steps in coaxial lines. The mathematics of the present work follows a similar pattern and therefore need not be given in full detail.

It may be shown that at a step discontinuity the potential difference between the two conductors is continuous, but that the current of the transmission line mode is discontinuous. It is therefore possible to represent faithfully the system, insofar as its behavior at terminals well away from the step is concerned, by a tandem arrangement of transmission lines with a shunt susceptance at the junction. This susceptance is such that the potential difference at the junction would drive through it a current equal to the jump in current at the step. It is in fact now well known that a step has the effect of a shunt capacitor.

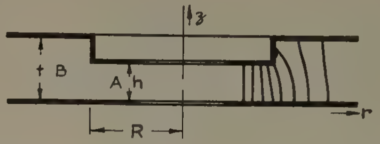

Fig. 2—Stepped disk transmission line—Case 1.

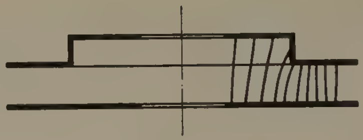

Fig. 3—Stepped disk transmission line—Case 2.

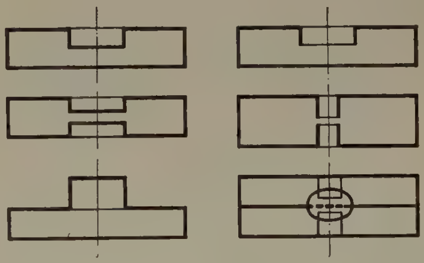

Fig. 4—Various forms of microwave apparatus embodying steps in the disk lines.

Fig. 5 shows the magnitude of the shunt capacity $C_d(\alpha)$ for steps in parallel plane transmission lines. We have

$$C_d(\alpha) = \frac{2\epsilon_0}{\pi^3} S_0(\alpha)$$

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† Formerly Div. of Radiophysics, CSIRO, Sydney, Australia; now Visiting Asst. Prof. of Radio Astronomy at the Univ. of California, Berkeley.

[1] N. Marcuvitz, "Principles of Microwave Circuits," McGraw-Hill Book Co., Inc. New York, N. Y., chap. 8; 1948.

[2] J. R. Whinnery and H. W. Jamieson, "Equivalent circuits for discontinuities in transmission lines," PROC. I.R.E., vol. 32, pp. 98–114; February, 1944.

[3] J. R. Whinnery, H. W. Jamieson, and T. E. Robbins, "Coaxial-line discontinuities," PROC. I.R.E., vol. 32, pp. 659–709; November, 1944.

where $C_d(\alpha)$ is the discontinuity capacity in picofarads[4] per centimeter run of step, α is the ratio of spacings, ϵ_0 is the permittivity of free space (0.08854 picofarads per centimeter) and $S_0(\alpha)$ is Hahn's function,[2,5]

$$S_0(\alpha) = \sum_1^\infty \frac{\sin^2 n\pi\alpha}{\alpha^2 n^3}.^6$$

Clearly if a disk transmission line filled with a medium of dielectric constant k has a step of amount α at radius R, then $2\pi RkC_d(\alpha)$ will be an approximation to the desired discontinuity capacity. The results of this paper are expressed as a correction factor to be applied to $2\pi RkC_d(\alpha)$. Where the correction factor does not depart significantly from unity, it is not necessary to go to the full theory, but the present work will be useful for determining the amount of error which the approximation involves. When the correction is important, it will be found to depend on whether the step-up is inwards or outwards.

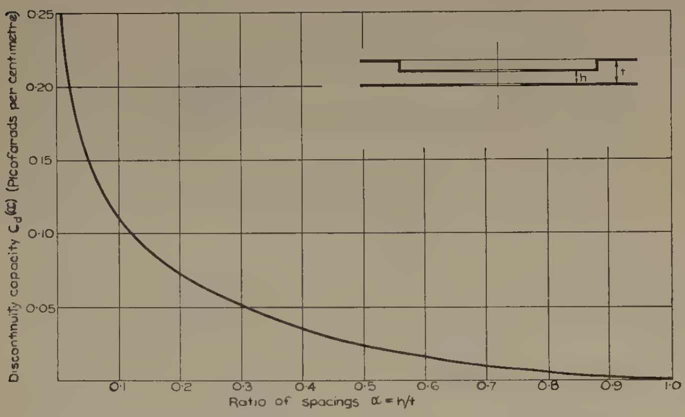

Fig. 5—Approximate discontinuity capacity per unit length of step.

Procedure for Using the Charts

Let it be required to calculate the discontinuity capacity for a step such as shown in Fig. 2. First, find the ratio of the spacings $\alpha = h/t$ and read off $C_d(\alpha)$ from Fig. 5. Multiply by $2\pi R$ the circumference of the step. This is the approximate discontinuity capacity in picofarads. Now find $\beta = R/h$, and from Fig. 6 read off $Bk(\alpha, \beta)$, the correction factor by which the approximate value must be multiplied.

When the step is as shown in Fig. 3, obtain the correction factor Bi from Fig. 7.

Readers whose interest is confined to using the charts for design will find the descriptions so far given sufficient for their purposes. For those interested in solving similar problems, the method is given below.

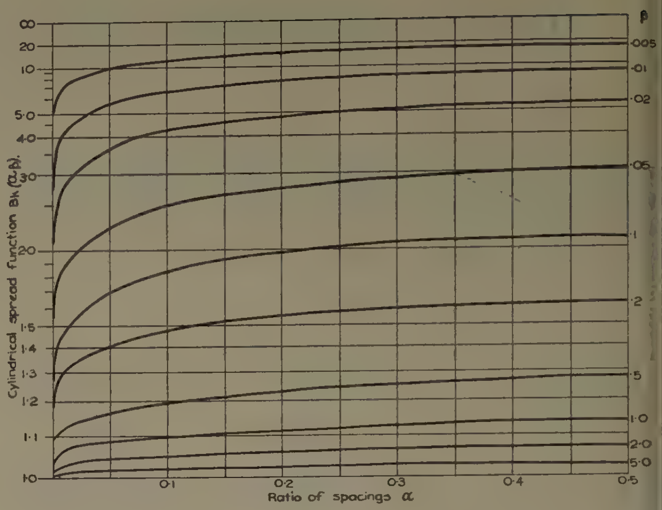

Fig. 6—Cylindrical spread correction factor $Bk(\alpha, \beta)$.

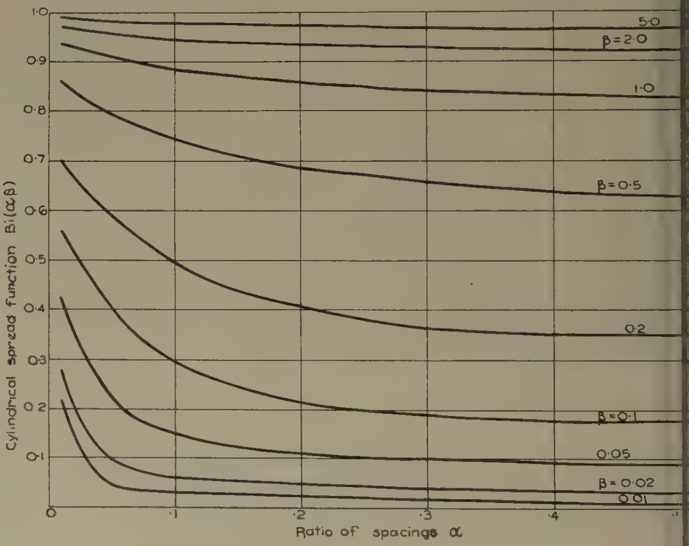

Fig. 7—Cylindrical spread correction factor $Bi(\alpha, \beta)$.

Theory

For the sake of simplicity, only the first case of the stepped disk transmission lines, that shown in Fig. 2, is dealt with, but the results for both are stated. Let a stepped disk line be divided into two regions A and B by a cylindrical surface containing the step. Then the field may be regarded as being that of the usual radial transmission mode, appropriate to each section of the line, together with a local superimposed field confined to the neighborhood of the gap, which allows the boundary condition to be fulfilled. It is assumed that the disks are close enough to prevent propagation of any but the principal wave, and that only circularly symmetrical modes are excited by the generator. The superimposed field is the sum of all circularly symmetrical modes, with axial components of electric field, which can exist between parallel conducting plates, each taken with proper phase and amplitude. To find these fields we write Maxwell's equations in cylindrical co-ordinates (r, θ, z) and specialize them by putting $\partial/\partial\theta = 0$ (circular symmetry), and as only radial and axial components of electric field will be required, we put $E_\theta = 0$. The magnetic field components H_r and H_z must then also be zero. This leaves three only of the original six partial differential equations, with the three simultaneous variables E_z, E_r, and H_θ. Solving these gives the required fields. It will be noted that the angular co-ordinate θ is not involved, that axial variations are described by circular functions, and that radial variations are described by the modified Bessel functions of zero order (I_0, K_0) and their derivatives (I_1, K_1). The local fields are of the form:

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[4] The term "picofarad" is the same as "micro-microfarad."

[5] W. C. Hahn, "A new method for the calculation of cavity resonators," *Jour. Appl. Phys.*, vol. 12, pp. 62–68; January, 1941.

[6] From work by Marcuvitz it may be shown that the expression $\pi^2\ln(4\alpha/e)$ is a useful approximation to $S_0(\alpha)$ for $\alpha \ll 1$.

$$E_z = \sum_{n=1}^{\infty} E_n \cos \frac{n\pi z}{t} \left[m_n I_0(P_n r) - K_0(P_n r) \right]$$

$$E_r = \sum_{n=1}^{\infty} \frac{n\pi}{P_n t} E_n \sin \frac{n\pi z}{t} \left[m_n I_1(P_n r) + K_1(P_n r) \right]$$

$$H_\theta = \sum_{n=1}^{\infty} \frac{-j\omega\epsilon}{P_n} E_n \cos \frac{n\pi z}{t} \left[m_n I_1(P_n r) + K_1(P_n r) \right]$$

where

$$P_n^2 = \left(\frac{n\pi}{t} \right)^2 - \omega^2 \epsilon \mu,$$

t is the separation of the planes, ω, ϵ, and μ have their usual meanings, and the E_n and m_n are constants of integration. For the appearance of the modified Bessel functions see Fig. 8.

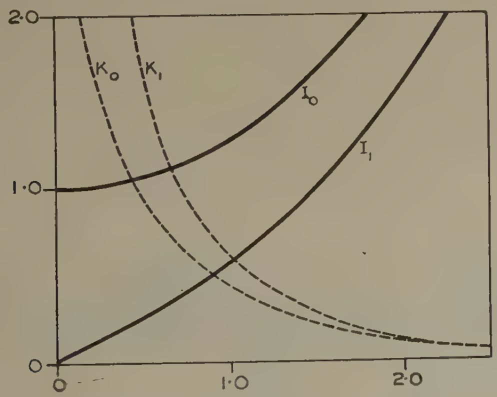


Fig. 8—The modified Bessel functions.

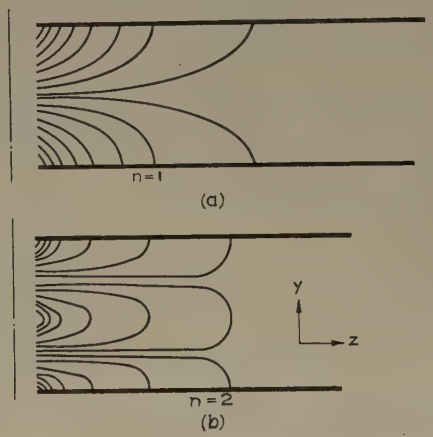
In region A of Fig. 2, only the I's are required and in region B only the K's, for the I's have a singularity at infinity and the K's have a singularity at the origin. Thus in region B, for example, the local axial electric field is

$$E_z = -\sum_{n=1}^{\infty} E_n \cos \frac{n\pi z}{t} K_0(P_n r).$$

Parts (a) and (b) of Fig. 9 show the electric field pattern corresponding to the first two terms of this infinite series. The component fields exhibit the familiar harmonic variation axially but quickly die away in the radial directions. This behavior is analogous to the exponential attenuation met with in uniform waveguides excited at a frequency lower than critical. However, the appropriate function in the present case must have a singularity at the origin. Moreover, the diminution with increasing argument must exceed that of the exponential law because of cylindrical spreading in two dimensions. The modified Bessel functions $K_0(z)$ and $K_1(z)$ have these properties. For large arguments both $K_0(z)$ and $K_1(z)$ have the asymptotic value

$$\sqrt{\frac{2}{\pi z}}\, \epsilon^{-z}.$$

Note the inverse square root factor which may be interpreted as attenuation due to cylindrical spreading, whilst the exponential core represents the usual attenuation occurring in cut-off uniform transmission lines and waveguides.



(a)

(b)

Fig. 9—(a) A partial field in region B. This field pattern corresponds to the first term in the series expansion of the local field. (b) A partial field in region B. The pattern described by second term.

Fig. 9(a) gives the field pattern of the first of the series of higher modes. It may be regarded as a radial transmission mode for which the conducting planes are too closely spaced to allow propagation. The critical spacing is the semi-wavelength of a plane wave in the medium. Still higher modes require even greater spacings before they can propagate and therefore, for a given spacing, attenuate more rapidly. Between plates one centimeter apart the attenuation "constant" for the first attenuated wave, if $\frac{1}{2}\lambda \gg 1$ (e.g. $\lambda = 3$ centimeters), falls away to an asymptotic value of about 3 nepers per centimeter at distances greater than about one centimeter from the axis. The attenuation is roughly proportional to the order of the wave and inversely proportional to the separation of the conductors. Fig. 10 (see following page) illustrates the general dependence of radial attenuation on the various parameters and allows it to be quickly determined in particular cases.

When we return to the system of Fig. 2, the problem is, with the partial fields at our disposal, to satisfy the boundary conditions at the radius of the step. On the surface of the post $h<z<t$ we have $E_z=0$ and E_r finite. In the gap $(0<z<h)$ we shall now assume $E_r=0$ and $E_z=$constant. It is not necessary to make this assumption, but in many cases the gap is so small that the field in the gap is distributed in a virtually pure E_{0m0} mode without axial variation. The assumption is equivalent to neglecting attenuated waves in region A.

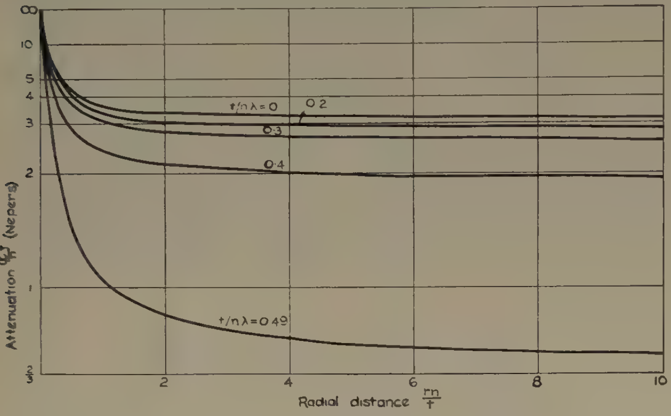


Fig. 10—Radial attenuation of higher modes in disk transmission lines. The attenuation "constant" α of the mode of order n, measured in units of t/n, is plotted against radial distance, also in t/n-units, for various ratios of t/n to the wavelength λ, where t is the separation of the conductors.

Let the axial field in the gap at radius R be $E(R)$ and let the field just outside the step be

$$E_0(R)+\sum_1^\infty E_n(R)\cos\frac{n\pi z}{t}\begin{cases}=E(R) & (0<z<h)\\ =0 & (h<z<t)\end{cases}$$

where $E_0(R)$ is the amplitude of the principal wave in region B at radius R, and $E_n(R)$ is the amplitude of the axial field of the nth attenuated wave in region B at radius R. Let the voltage across the gap be

$$V=E(R)h.$$

Then this voltage appears in full across the mouth of disk line B, because the attenuated fields contribute nothing to the line integral of electric field across the cavity. Thus when $n\neq 0$

$$\int_0^t E_n(R)\cos\frac{n\pi z}{t}\,dz=0.$$

Hence

$$\int_0^t\left[E_0(R)+\sum_1^\infty E_n(R)\cos\frac{n\pi z}{t}\right]dz=E_0(R)t$$
$$=E(R)h=V.$$

As the voltage of the principal waves in the two regions is thus continuous across the step, it follows that, any change in input reactance to the outer line due to the presence of the discontinuity, will appear as a difference between the current leaving region A, and the curren of the principal wave in region B. The difference curren may be taken to flow through a susceptance shunte across the junction of the disk transmission lines.

Let the magnetic field just outside the step be

$$H(R)=H_0(R)+\sum_1^\infty H_n(R)\cos\frac{n\pi z}{t}$$

where $H_0(R)$ is the field of the principal wave in regio B. Let the field just inside the step be

$$H_A(R)=H_{0,A}(R)+\sum_1^\infty H_{n,A}(R)\cos\frac{n\pi z}{h}\cdot$$

Since the fields must match at the gap,

$$H_A(R)=H(R),\qquad (0<z<h)$$

If we expand each side of this equation in Fourier serie of interval h, it follows that the field of the principa wave in region A, $H_{0,A}(R)$ is given by the average o $H(R)$ taken over the gap:

$$H_{0,A}(R)=\frac{1}{h}\int_0^h H(R)dz$$
$$=H_0(R)+\frac{1}{h}\int_0^h\sum_1^\infty H_n(R)\cos\frac{n\pi z}{t}\,dz.$$

The discontinuity in current has the magnitude

$$2\pi R[H_{0,A}(R)-H_0(R)]=\frac{2\pi R}{h}\int_0^h\sum_1^\infty H_n(R)\cos\frac{n\pi z}{t}\,dz$$
$$=\frac{2\pi R}{h}\sum_1^\infty\frac{t}{n\pi}H_n(R)\sin\frac{n\pi h}{t}\cdot$$

Hence the shunt susceptance is given by

$$jB=\frac{2\pi R[H_{0,A}(R)-H_0(R)]}{E_0(R)t}\cdot$$

It now remains to make the proper substitution fo $H_n(R)/E_0(R)$ the relative amplitude of the highe modes with respect to the field of the principal wave Omitting the proof, the result is

$$\frac{H_n(R)}{E_0(R)}=\frac{j\omega\epsilon K_1(P_nR)2t}{P_nK_0(P_nR)n\pi h}\sin\frac{n\pi h}{t}$$

and hence for the shunt capacity

$$C=\frac{4Rt\epsilon}{\pi h^2}\sum_1^\infty\frac{K_1(P_nR)\sin^2\dfrac{n\pi h}{t}}{K_0(P_nR)n^2P_n}\cdot$$

This formula is suitable for computing. If $t\ll\lambda$ s that we may take $P_n=n\pi/t$

$$C = \frac{4R\epsilon}{\pi^2}\left(\frac{t}{h}\right)^2 \sum_1^\infty \frac{\dfrac{K_1(n\pi\beta)}{K_0(n\pi\beta)}\sin^2 n\pi\alpha}{n^3}$$

where

$$\alpha = \frac{h}{t}$$

$$\beta = \frac{R}{t}.$$

We may rewrite this result

$$C = \frac{4R\epsilon}{\pi^2} C(\alpha, \beta).$$

Since the property of the C-function is that

$$\lim_{\beta \to \infty} C(\alpha, \beta) = S_0(\alpha)$$

where $S_0(\alpha)$ is Hahn's function of zero order, we normalize $C(\alpha, \beta)$ with respect to $S_0(\alpha)$ and write

$$C = \frac{4R\epsilon}{\pi^2} S_0(\alpha) Bk(\alpha, \beta)$$

and the new function $Bk(\alpha, \beta)$ is defined by

$$Bk(\alpha, \beta) = \frac{\displaystyle\sum_1^\infty \frac{K_1(n\pi\beta)}{K_0(n\pi\beta)} \frac{\sin^2 n\pi\alpha}{\alpha^2 n^3}}{S_0(\alpha)}.$$

A chart of this function is given in Fig. 6.

There is a corresponding function which arises from the case illustrated in Fig. 3 and is defined by

$$Bi(\alpha, \beta) = \frac{\displaystyle\sum_1^\infty \frac{I_1(n\pi\beta)}{I_0(n\pi\beta)} \frac{\sin^2 n\pi\alpha}{\alpha^2 n^3}}{S_0(\alpha)}.$$

This function replaces $Bk(\alpha, \beta)$ in formulas relating to the second sort of stepped line. A chart is given in Fig. 7.

Now since

$$C = 2\pi R k C_d(\alpha) Bk(\alpha, \beta),$$

$Bk(\alpha, \beta)$ may be regarded as a curvature-correction factor to take account of the curvature of the step. It will be noticed that $Bk(\alpha, \beta)$ and $Bi(\alpha, \beta)$ lie on opposite sides of unity. Consequently the capacity is greater in the one case, and less in the other, than that calculated on the parallel plane assumption.

CORRECTION

Laurence G. Cowles, author of the paper, "The Parallel-T Resistance-Capacitance Network," which appeared on pages 1712–1717 of the December, 1952 issue of the PROCEEDINGS OF THE I.R.E., has brought the following correction to the attention of the editors:

In Eq. 6 on page 1713, the right side should be multiplied by $\frac{1}{2}$, and r should be inserted in the last term of the denominator, to correct an obvious dimensional error.

Positive-Ion Trapping in Electron Beams*

E. L. GINZTON†, FELLOW, IRE, AND B. H. WADIA†, ASSOCIATE, IRE

Summary—Residual gas within the envelope of an electron tube forms positive ions by electron collision. Ordinarily these ions are quickly lost to the cathode; however, if they can be held within the beam space, the spreading due to the space charge of the beam might be considerably reduced. A "trapping" electrode, with an appropriate positive potential, placed at the cathode end of the drift tube was believed sufficient to affect trapping. Since this simple method was first proposed, attempts to make the scheme work yielded inexplicable results. A detailed study of the phenomenon reveals that the trapping process is more complex than was originally believed. A qualitative theory is advanced which is consistent with experiment and leads to suggestions for an improved performance. By a proper design of the trapping electrode, transmission through a drift space was improved from twelve per cent to eighty per cent.

I. INTRODUCTION

THE EFFECT OF space charge in an electron stream is to produce forces which tend to increase its cross section. Such beam spreading is often serious, and must be counteracted by some means. The common method in use today (in klystrons and traveling-wave tubes) is to subject the electron flow to an axial-magnetic focusing field.

A simpler method, which has been known for some time,[1] consists of utilizing the unavoidable positive ions to neutralize the negative space charge of the electrons and thus to suppress the cause of the divergent forces. The positive ions are created within the electron stream by collision of the electrons with molecules of residual gas. There is a tendency for the ions so formed to migrate towards the axis of the electron stream until it is over-neutralized and becomes positive. Such a neutralized beam will then lose ions in several ways, the important one being the flow of positive ions towards the cathode. It has been suggested previously that if this loss towards the cathode could be eliminated by interposing a suitable potential rise in the path of the ions, the electron stream would by necessity become over-neutralized even at the very best vacuum.

Such a rise in potential can be introduced by means of an electrode (placed near the cathode end of the drift space) which is concentric with the beam. This electrode may be in the form of a grid or may consist of a gridless structure. Gridded type of trapping electrode will not be considered in this article because of several reasons; first, considerations of power dissipation and noise often preclude the use of a grid, and secondly, the theory of the gridless trap is general and includes the simpler case of a grid stretched across the beam.

Through private correspondence, we have knowledge of a few attempts to make use of gridless positive-ion traps; these, apparently, have not met with complete success. In fact, they have yielded results which were not explicable on the basis of the simple theory previously available.

It was, therefore, our desire to understand more completely the mechanism of ion neutralization in *gridless* ion-trapped electron streams. We have found it possible to extend and modify the simple theory of Field *et al.*,[2] and to obtain a nearly complete understanding of the resulting phenomenon. Several experiments conducted during these studies agree with the theory which will be presented below. As will be shown, it is certainly possible to provide positive-ion traps which can assure complete neutralization of electron beams; but, as will also appear below, such ion traps are not universally useful. It will be found that it is extremely difficult to provide ion traps for beams that nearly fill the metallic tube through which the beam may be passing, but that it is relatively easy to provide such traps if the diameter of the beam is small in comparison with the diameter of the metallic walls.

II. THE UNTRAPPED POSITIVE IONS

A. Neutralizing Effect of Ions

At the generally accepted good vacuums, a large number of gas molecules is usually present within the tube envelope; the presence of these molecules results in the formation of positive ions by virtue of their collisions with electrons. The rate of positive-ion formation depends on the nature of the gas, its pressure, the number of electrons present in the beam, and their velocity.

Since these ions are positively charged, they will move toward points of lower potential, near the axis of the beam. The slow electrons, which are the other product of the ionizing collision, will move in the opposite direction (to the drift tube), and will be lost. The beam will thus form a core of positive ions clustered around its axis. If, for the present, we assume that only few ions are lost, a sufficient amount of positive charge must eventually collect within the beam to neutralize the electron charge completely.

Any excess ions that are created constitute a net positive charge that results in a radial "hill" of potential down which these surplus positive charges are drained to the drift tube. Hence, under the assumption of no loss, the number of ions in the beam must closely approach the number of electrons if a sufficient amount of gas is available for ionization.

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* Decimal classification: R138.1. Original manuscript received by the IRE, June 3, 1954. The research reported in this article was carried out as a project supported by Joint Services Contract N6 251, T.O.7. A more complete treatment appeared in Technical Report No. 74, "Influence of positive ions on electron-beam profiles," by B. H. Wadia, published at Stanford University (Electronics Research Laboratory) on January 28, 1954, and in a doctoral dissertation submitted to the Department of Electrical Engineering, Stanford University. This material was also presented as a paper before the 1954 IRE National Convention, New York, N. Y., March 25, 1954.

† Stanford University, Stanford, California.

[1] L. M. Field, K. R. Spangenberg, and R. Helm, "Control of electron-beam dispersion at high vacuum by ions," *Elect. Comm.*, vol. 24, pp. 108–121; 1947.

[2] *Ibid.*

B. *Minor Causes of Ion Loss*

The assumption that no ions are lost from the beam space is, of course, not justified. Several processes which cause a continual removal of ions from the drift tube may be listed and their effects may be shown to be negligible by consideration of their magnitudes:[3]

1. Ions and electrons recombine within the beam to form neutral molecules. However, at pressures in the vicinity of 10^{-7} mm Hg, the rate of ion formation far exceeds the rate of recombination.

2. The thermal energy of the ions is of the order of kT, corresponding to about 0.025v. This is negligible when compared to the normal potential variations across the beam which tend to hold the ions together. The loss due to thermal energy might become important as one approaches the state of complete neutralization because, under such conditions, the potential variation across the beam becomes vanishingly small.

3. The impact energy given to the ion by the colliding electron is small when compared to the potential drop in the beam. Like the loss due to the thermal energy, this cause of ion removal will not assume serious proportions until a nearly neutralized condition is reached.

4. An insulator in the vicinity of the beam becomes negatively charged by picking up stray electrons and hence constitutes a sink for positive ions. This can be very serious; however, most electron beams can be surrounded partially or completely by metallic structures and can, therefore, be shielded from the effects of the sinks described above.

C. *The Major Cause of Ion Loss*

A more careful consideration of the above discussion would make it seem that electron beams enclosed in metallic drift spaces should reach total neutralization even at pressures as low as 10^{-7} mm Hg since, at such pressures, a sufficient number of gas molecules is still available to form the required number of ions.

That this is not the case was shown by Field, *et al.*,[4] who described a major source of ion loss that had not been considered previously in the literature. The accelerating field from the cathode-anode region penetrates into the drift space via the anode aperture through which the beam must enter. This field is strongest at the axis near the aperture and is of such a polarity as to remove ions there. The removal of the ions at the head of the beam leaves a low-potential region to which ions must flow from parts of the beam further down the drift space. Thus, the effect of the draining field is felt for a substantial distance along the beam.

In fact, it can be shown[5] that only a slight amount of neutralization is possible when pressures lower than 10^{-7} mm Hg are considered. Therefore, the spreading of an electron beam, under the pressure conditions normally available in commercial tubes, may be taken to be practically the same as in perfect vacuum.

III. The Elementary Theory of Ion Trapping

The principle first introduced by Field, which helps to eliminate axial drainage, may be explained by considering Fig. 1. A thin beam of paraxial electrons is flowing through a comparatively large drift space as shown. Let it be our intention to obtain a completely neutralized beam at all points to the right of section AA. Some means is then necessary to prevent the positive ions from flowing across section AA toward the cathode. When this is done, all the ions created in the beam to the right of AA will tend to be held within the beam boundaries. The removal of ions can then be caused only by the minor processes mentioned previously; the rate of removal due to these is so small that a sufficient number of ions should collect to neutralize the beam completely, even at very low pressures.

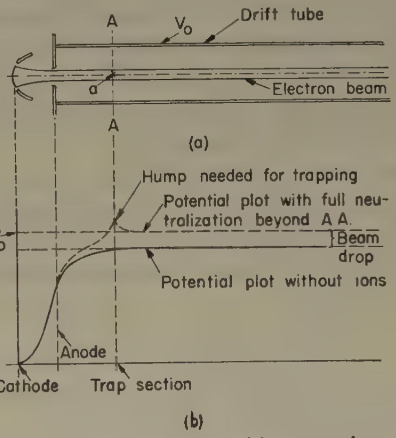

Fig. 1—Principle of axial ion trapping.

To understand the stopping of ion flow across a given section, examine the potential distribution in that vicinity. On the left-hand side of AA, the positive ions are drawn rapidly toward the cathode so as to leave that part of the beam unneutralized. The potential on the axis there will always be less than V_0, the drift-tube voltage, by some amount which depends upon the parameters of the beam. If the beam holds ions to the right of AA, the potential on the axis to the right of AA approaches V_0 to an extent dependent upon the degree of neutralization. For the case of "full" neutralization, all points to the right of AA should reach the potential V_0, since no net negative charge remains.

In order to retain ions in this way, the potential on the axis at a must be raised above the value V_0, so that a hump of potential is interposed in the path of the ions, as shown in Fig. 1(b).

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[3] L. M. Field, "Theory and measurement of electron beam dispersion phenomena at high vacuums," Stanford University Ph.D. Thesis; March, 1944.

[4] L. M. Field, K. R. Spangenberg, and R. Helm, *op. cit.*

[5] B. H. Wadia, "Influence of positive ions on electron-beam profiles," Stanford University Ph.D. Thesis; February, 1954.

A grid held just above the drift-tube potential is the simplest means of obtaining the potential hump. It will become clear subsequently that this is the only method which can give perfect ion trapping. However, considerations of power dissipation often preclude the use of suitable grids. Aperture plates, cylinders, and ring electrodes concentric with the beam have been suggested as alternative means of obtaining the required variation of potential along the axis. Such *gridless ion trapping* is the subject of the present discussion. As will be shown, many complications arise when this method of trapping is used; as a result, the design of such a system of beam control becomes extremely complex. The simple theory outlined above needs to be extended and modified in order to explain the behavior of *gridless* ion traps.

IV. Experimental Data on Gridless Ion Trapping

The basic theory was confirmed by Field by means of the following experiment. He produced an initially parallel beam at 7300v and 130 ma. This beam would normally spread to twice its starting diameter in a distance of 36 cm. By providing a trapping aperture at the entrance, a 15-v bias was found to be sufficient to accomplish complete neutralization; and such neutralization was found to prevent the beam spreading. An increase in trapping voltage did not affect the behavior of the beam appreciably.

Attempts to make use of the ion trap as described by Field have been made by several groups, but without any success.[6] Additional experiments clearly indicated that something was missing from the idealized picture presented in Field's concept of ion trapping.

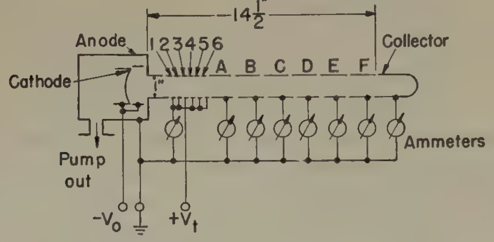

Fig. 2—Beam-tester schematic diagram.

It was, therefore, decided to perform an experiment to study more carefully the behavior of a beam controlled by a gridless ion trap and operating at voltages and currents of interest in contemporary klystrons and traveling-wave tubes. The schematic diagram of Fig. 2 shows the essential parts of the equipment used in our experiment. The beam-testing tube[7] consists of a bombarded tantalum cathode injecting a beam of perveance 1.07×10^{-6} and an entrance angle of 2.5 degrees with an initial diameter of 2.54 cm into a drift tube of the same diameter. Its anode is followed by ring electrodes, numbered one to six, to be used for trapping. The position of the trapping section may be varied by suitably biasing one out of the six electrodes. The rest of the drift tube is broken up into six sections and a collector, insulated from each other by glass seals.

Curves of the observed percentage of total current reaching the collector as a function of the trapping voltage for different positions of the trapping section are given in Fig. 3. A glance at these curves indicates a definite pattern of behavior which cannot be explained by the simple ion-trapping theory.

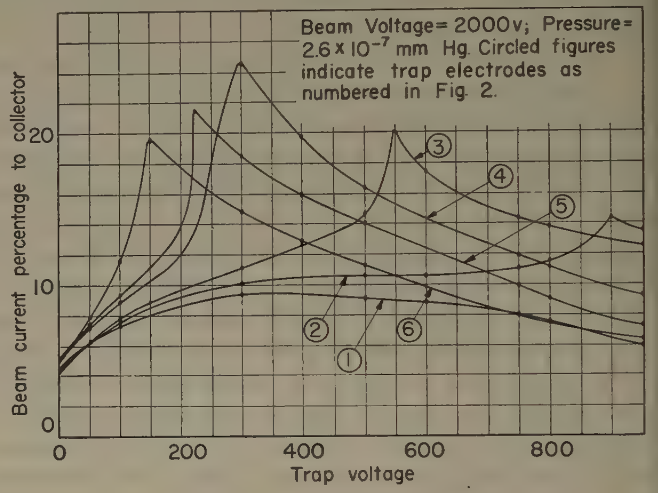

Fig. 3—Percentage beam transmission as a function of the trapping voltage.

It is to be observed that the current transmission improves in all cases with application of trapping voltage, and that the transmission is best when the trapping is applied to bias ring No. 4—the place where the beam should reach its minimum diameter as predicted by the gun design; but, even at best, the transmission through the drift tube is only 25 per cent—far short of the ideal transmission that could be expected from a perfectly neutralized beam.

V. Detailed Consideration of Gridless Trapping

The process of ion trapping as discussed in section III is assumed to be a purely axial phenomenon; the movement of the ions and the insertion of the potential hump are considered as being along the axis. While these assumptions may hold for thin beams of low perveance, a more general theory should consider trapping as a volume phenomenon. Therefore, one needs to look more closely at the fields involved in the region around the trap and to study the effect which these fields have on the distribution of ions.

A. Potential Distribution Around the Trapping Electrode

The potential in the drift tube is the result of contributions from three different causes. First, the positive voltage on the trapping electrode raises the potential at every point in the drift space. Secondly, the presence of the negative particles in the beam tends to lower it. Thirdly, the *combined* effect of these two causes results in trapping positive ions, the presence of which will modify the potential due to the other two, depending on the particular distribution of the ions.

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[6] Private communication with workers at Varian Associates, San Carlos and Palo Alto, California.

[7] The construction and processing of this beam-testing tube was made possible through the co-operation of Eitel-McCullough, Inc., San Bruno, California.

The potential distribution due to the trapping-electrode voltage V_t (shown dotted in Fig. 4) can be obtained by electrolytic-tank measurements. The distribution due to the beam may be obtained approximately by solving the problem of a cylinder of charge coaxial with a conducting wall. Since potentials are additive, these two may be added numerically (or graphically) in order to obtain the resultant potential distribution.

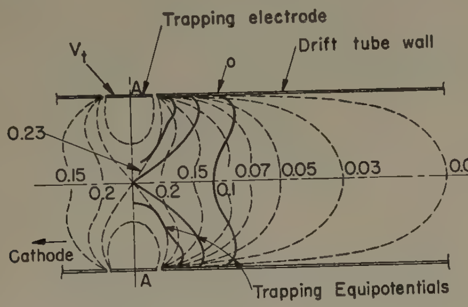


Fig. 4—Potential distribution near the trapping electrode.

It will soon be clear that the only equipotential contour relevant to our discussion is the one having the same potential as the drift tube. Such an equipotential must exist, because there are points within the drift space where the increase in potential due to the positive trapping voltage just cancels the drop due to the negative space charge of the beam. The net potential at these points, in the absence of ions, will be equal to the drift-tube potential.

The location of this contour line depends on the value of the trap voltage and the parameters of the electron beam being used; a separate set of calculations must be made for each set of conditions. Figs. 4 and 5 show contours calculated for three values of trap voltage and a 2000-v beam of perveance 1.07×10^{-6}, just filling the drift tube.

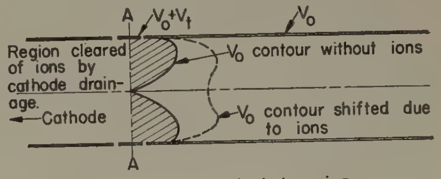
The contribution made to the potential distribution by the presence of positive ions will alter the location of the V_0 contour; but the position of this contour, as will be seen subsequently, determines the distribution of positive ions. This interdependence makes it impossible to determine the actual distribution of the ions and the final position of the V_0 equipotential. Consequently, the rest of the discussion is based on qualitative arguments.

B. The Trapping Phenomenon

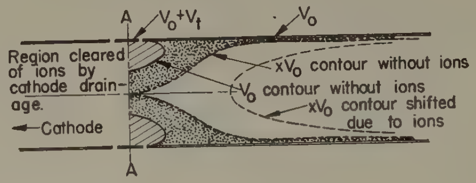
At all points in the beam, positive ions are continuously being formed; their potential energies will depend on the value of the potential at the place where they are created. Ions that are formed at a place where the potential is high with respect to its surroundings will tend to flow away from the point in much the same way as water flows away from tops of hills. The rate of formation of ions at pressures of the order of 10^{-7} mm Hg is so slow that, under such pressure conditions, the rate of ion drainage due to even the smallest fields can exceed the rate of formation; high-potential regions, such as the one mentioned above, may be taken to be virtually free of ions. Conversely, ions formed in places that are surrounded completely by a surface of higher potential cannot have sufficient energy to pass this bounding surface, and hence they must remain trapped therein.

The distribution of ions within a given beam space therefore depends on the nature of the potential contours within that space. In order to get a qualitative idea of the disposition of the ions at different values of trapping voltage, let us consider the potential contours shown for the three cases in Fig. 5.

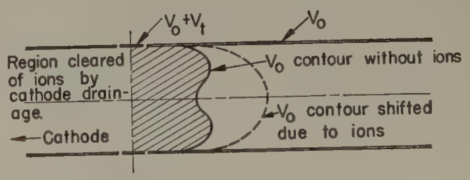
Fig. 5(a) shows the contour line which should, by the simple axial theory of section III, have trapped enough ions completely to neutralize the beam to the right of the trap. The portion of the diagram shown shaded is a region where the potential is higher than V_0; hence, all the positive ions formed in that region are drained away either to the walls or to regions of lower potential. The shaded region cannot, therefore, have its space charge neutralized to any appreciable degree.


A
V_0+V_t
V_0
Region cleared of ions by cathode drainage.
Cathode
V_0 contour without ions
V_0 contour shifted due to ions
A

(a) V_t adjusted to just start trapping


A
V_0+V_t
V_0
Region cleared of ions by cathode drainage.
Cathode
xV_0 contour without ions
V_0 contour without ions
xV_0 contour shifted due to ions
A

(b) Low value of V_t


V_0+V_t
V_0
Region cleared of ions by cathode drainage.
Cathode
V_0 contour without ions
V_0 contour shifted due to ions

(c) High value of V_t

Fig. 5—Movement of the trapping surface as the trapping potential is varied.

The space that is left unshaded is essentially one of potential below V_0, and is also a region which is completely surrounded by the V_0 equipotential; hence it is isolated from the draining field of the cathode. Positive ions found in this space will remain there and begin to neutralize the electron beam. As soon as neutralization starts, the V_0 equipotential begins to shift toward the right because of the accumulation of positive charge to the right of it. This shift makes the trapping region move down the drift tube as positive ions become trapped. Ions will continue to be trapped until the whole space enclosed by the new V_0 equipotential has reached the drift tube voltage. After this, the positive charges will begin to move out to the drift tube and be lost.

Fig. 5(b) shows the potential distribution when the value of V_t is below the minimum prescribed by the simplified theory of section III. The space shown cross-hatched is a region with potentials above V_0 and hence must be essentially free of ions. The area marked with dots is below V_0 in potential, but it has a direct connection with the unneutralized low potential part of the drift tube to the left of section AA where the cathode is continually removing ions. This region, therefore, cannot retain ions.

Trapping can only take place inside a space enclosed completely by some such potential contour as is shown dotted-and-dashed in Fig. 5(b). Let this potential be denoted by xV_0, where $x<1$. As before, when trapping begins, the xV_0 equipotential will begin to shrink because of the positive space charge enclosed within it. Some final equilibrium distribution will be established, consistent with the above considerations. Under such conditions, the neutralization cannot be very great: first, because the trapping space is small, and secondly, because trapping takes place within a contour of potential lower than V_0. This is analogous to the fact that a small shallow bowl holds less water than a deep bowl does.

Fig. 5(c) illustrates the potential distribution for high values of V_t. A little reasoning will show that, the higher one takes V_t, the farther to the right does the V_0 equipotential shift, thus increasing the size of the unneutralized portion of the beam.

The treatment of the three cases above gives an idea of the disposition of the trapping zone with respect to the trap electrode for different values of voltage. It must not be imagined, however, that ions attain a uniform density at all points inside the trapping region. Their behavior in this space is more a matter of conjecture than of accurate mathematical calculation. The phenomenon might, perhaps, be more clearly understood if one were to liken the trapping zone to a sunken area of ground being filled up by a slow seepage of water. The deepest portions of the hollow will, of course, be filled first and eventually will hold the largest depths of water, while the surrounding areas near the lip of the hollow will be filled last and will constitute the shallow region. In the electrostatic case, matters are further complicated by the effect which the positive ions themselves have in altering the potential distribution in the surrounding domain. Despite this complication, the same general reasoning holds, and the qualitative picture remains the same.

C. *Effect on Beam Transmission*

It has been established above that, for a given trapping voltage, there exists within the drift tube a surface that encloses a certain volume of the drift space inside which positive ions are held. The positive ions will accumulate rapidly in that part of the beam, until the whole volume enclosed by the surface mentioned is practically at uniform potential, after which the excess positive ions will migrate to the drift-tube wall.

An electron entering this neutralized space will be entering an equipotential region and must therefore maintain the magnitude and direction of the velocity with which it enters. The slope of the trajectory of each electron as it crosses into the trapping region is hence a parameter of great importance in determining the subsequent behavior of the beam.

1. The Perfect Ion Trap. It would be best to start out by examining what is needed in order to obtain a truly parallel beam by means of ion trapping. In view of what has been said above, the principle of focusing by trapping positive ions may be stated as follows: *The various electrons comprising the beam must be made to enter the unipotential region when their paths have zero slope.* If one can attain this condition, all the electrons of the beam must continue to travel from there on with zero slope, since no radial forces exist in a unipotential region. Perfect transmission should, therefore, be obtained if the minima of all the electron trajectories are made to coincide with the position of the trapping surface.

2. Realization of the Ideal Conditions. In order to understand the extent to which perfect trapping may be achieved, the following two points must be considered:

(a) The contour behind which ions are trapped is some surface of revolution (similar to the ones shown dashed in Fig. 5) the existence of which has been established but the actual shape and position of which cannot be determined by analytical, graphical or numerical methods.

(b) In order that the minima of the trajectories of the beam electrons may be made to coincide with the position of the trapping surface, the location of these minima should be accurately known. The most complete treatment of this problem available to date[8] suffers from two drawbacks. First, an estimate of the position of the beam minimum obtained from this treatment is only an approximate one, due to the assumptions inherent in setting up the problem. Secondly, the treatment only deals with the approximate behavior of the electron at the edge of the beam; no mention is made of the behavior of the electrons within the beam envelope. The reduction of electron velocities which accompanies the drop in potential in an unneutralized beam will cause an accumulation of charge in the regions where the velocity is reduced, i.e., the charge density will be higher near the axis than at the edge of the beam. The inner electrons will be subject to greater repelling forces since their trajectories enclose a larger average charge density, and hence will go through their minima quicker than the outer electrons; therefore, the minima will not all lie in the same cross-sectional plane.

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[8] L. P. Smith and P. L. Hartman, "Formation and maintenance of electron and ion beams," *Jour. Appl. Phys.*, vol. 11, pp. 220–224; 1940.

Fig. 6 shows a sketch of the relative positions of the trapping equipotential and the beam minima as they might possibly occur at some stage during the trapping experiment. It is also shown in the figure that the different electron trajectories enter the neutralized zone at different slopes and hence behave differently subsequent to their entry. Therefore, while electrons at a particular radius will remain parallel to the axis, the rest of the electron paths are at different slopes. Crossovers could very well occur and complete transmission is improbable.

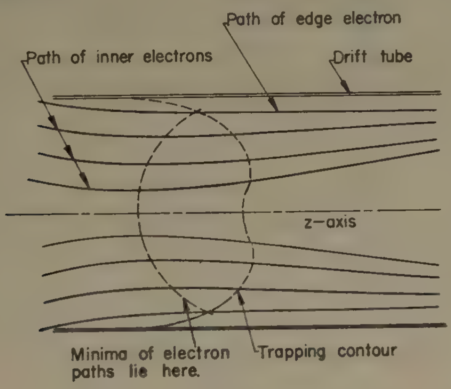

Fig. 6—The trapping contour and the positions of the minima.

VI. Explanation of the Experimental Results

Inasmuch as a strict mathematical treatment of the trapping phenomenon is not possible, quantitative estimates cannot be made to match with the values of transmission obtained by experiment. But the qualitative argument of section V does indicate a definite trend of behavior which seems to be borne out by our experiment.

A. *Effect of Trapping Voltage*

As the trapping voltage is increased from zero, a rapid rise in beam transmission is first observed, followed by a slower fall. This is not entirely unexpected when one considers the phenomena outlined in connection with Fig. 5.

At really low trapping voltages, the draining effect of the cathode field can still leak into the drift tube beyond the trapping ring, and only partial neutralization can exist as was explained in connection with Fig. 5(b). As one increases the trapping voltage, the larger value of V_t offers better opposition to the cathode drainage, and transmission improves.

However, excessive trapping voltages have the effect of pushing the neutralized region down the drift space. The slope of the beam under such circumstances would be positive as it enters the trapping region, thus resulting in a diverging beam.

Between these two effects, there results some "optimum" condition where the transmission reaches its greatest value before beginning to fall. Such a peaked behavior is shown by each of the experimental curves.

B. *Position of Trap*

It is interesting to observe the effect of shifting the position of the trapping electrode. It is seen that, as one moves the trapping section closer to the cathode, the voltage V_t required to achieve maximum transmission becomes higher. This is probably because as one gets closer to the cathode its draining effect becomes greater and one needs larger trapping voltages to counteract the unwanted field.

Another effect seen from the family of experimental curves in Fig. 3 is that the value of maximum transmission obtained is different for different positions of the trapping ring. One may refer to section V-C-2 in order to find the explanation for this effect. It was pointed out there that the position of the trapping surface relative to the beam minimum has a considerable effect on the resultant transmission. If the trapping surface falls in the region where the beam diverges, the electrons will continue with the same positive slope and only a slight improvement can be expected. When the trap voltage is applied to the electrode close to the cathode, the beam is still converging as it enters the trapping zone and should continue to converge until it crosses over and then diverges. The position of the trapping ring which brings the trapping surface closest to the minimum will give the best possible results.

VII. Conclusions

It has been found possible to obtain an improved understanding of the process of gridless ion trapping by considering it as a volume phenomenon. A detailed examination reveals that this method of controlling beam spreading involves more complications than are accounted for in the simple theory (which theory is only applicable in the special case of thin beams of low perveance).

The conditions for perfect trapping can be simply stated; however, it is not very easy to achieve them with gridless traps. The main difficulty lies in the fact that the designer cannot predict the distribution of the positive charges within the beam; as a result, the position of the unipotential region (formed by the accumulation of positive ions) is no longer a matter of deliberate design. Hence, it is not possible to obtain a trap configuration which will assure that the various electrons in the beam will enter the trapping region precisely at the moment when the slopes of their trajectories are zero.

There have been instances of efficient trapping with gridded drift tubes in some experimental klystrons. This is certainly understandable in view of what has been discussed. By positioning the grid at the estimated minimum of the beam, the trapping equipotential can be forced to stay close to the minima of the electron trajectories.

In the absence of a proper analytical design, several gridless electrode configurations can be tried out experimentally. Chances of success should be better in the case of those electrode configurations having potential distributions that come close to that of a grid (at drift-tube potential) stretched across the beam at the point of its minimum radius. One such example would be that of an Einzel lens[9] around the minimum of a beam the diameter of which is small compared to the drift space.

Using the arguments discussed above as a guide, a tube was constructed as shown in Fig. 7. A beam of perveance 2.6×10^{-6} and a minimum diameter of 5/32 inch was injected into a drift tube of 0.5 inch diameter and 6 inches long. For a beam voltage of 300v, over 80 per cent of the cathode current was received at the collector when 95v was applied to a trapping electrode of the Einzel-lens type situated near the minimum of the beam.

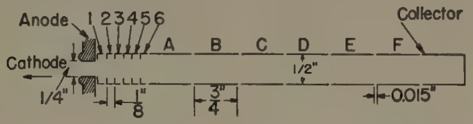

Fig. 7—An experimental tube to test an improved trap design.

If it is desired to make an electron beam nearly fill a metallic drift tube (as is usually required in a klystron), a constriction in tube diameter can be provided at some distance behind the gridless ion trap, as shown in Fig. 8. Thus, it is possible to achieve a configuration suitable for ion trapping and still have a metallic tube "hug" the beam as is needed in rf interaction regions.

It is not unreasonable to expect that subsequent trials with different types of trapping devices could yield further improvements in beam transmission.

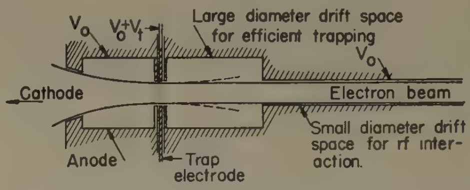
V_0
V_0+V_t
Large diameter drift space for efficient trapping
V_0
Cathode
Electron beam
Small diameter drift space for rf interaction.
Anode
Trap electrode

Fig. 8—Trap design for a tube which requires that the beam fill the rf interaction space.

VIII. Acknowledgment

The authors are greatly indebted to Eitel-McCullough, Inc., of San Bruno, California for their aid in constructing the beam-tester, and to Dr. Charles Susskind, Stanford University, Stanford, California for assistance in the preparation of this paper.

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[9] K. R. Spangenberg, "Vacuum Tubes," McGraw-Hill Book Co., Inc., New York, N. Y., p. 386; 1949.

Impedance Measurement by Means of a Broadband Circular-Polarization Coupler*

SEYMOUR B. COHN†, SENIOR MEMBER, IRE

***Summary*—This paper describes a waveguide component particularly useful as a broadband impedance-measurement instrument. The component consists of a length of circular waveguide at right angles to the broad wall of a rectangular waveguide, and coupled to it by three apertures. A probe rotated one turn in the circular waveguide will sample a field variation exactly equivalent to that in one guide wavelength of the rectangular waveguide. This condition may be maintained with high accuracy over approximately a 20-per cent bandwidth. The principle of operation of the device is discussed in the paper and design equations given. Several possible applications are described.**

Introduction

THE BROADBAND waveguide coupling device described is particularly suitable for use in automatic or manual impedance-measurement apparatus. The device consists of a rectangular waveguide coupled by an arrangement of three apertures to a circular waveguide joined at right angles to the broad wall of the first waveguide (Fig. 1). A TE_{10} wave traveling from left-to-right in the rectangular waveguide will excite a circularly polarized TE_{11} wave in the circular waveguide. Similarly, a wave traveling from right-to-left will also excite a circularly polarized wave, but one having the opposite rotation. By a proper co-ordination of three apertures, it is possible to maintain this circular polarization with high accuracy over approximately a 20-per cent frequency band. Similar coupler designs using only one aperture have previously been proposed,[1,2] but are limited to a bandwidth of about two per cent. These earlier designs, therefore, have had little use.

The basic property of this coupler that gives it value as an impedance-measurement instrument is its ability to reproduce a standing wave in the rectangular waveguide as a field-strength variation around the circumference of the circular waveguide. This varying field, which may be sampled by a probe rotated in a given transverse plane about the axis of the circular waveguide, has a maximum-to-minimum ratio equal to the vswr in the rectangular waveguide. A second important property of the coupler is that the angular position of the minimum varies linearly with the position of the minimum in the rectangular waveguide, with the former in mechanical degrees being equal to the latter in electrical degrees measured from a suitable reference plane. In other words, one rotation in the circular guide is exactly equivalent to one rectangular-waveguide wavelength at all frequencies in the operating band.

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* Decimal classification: R244.21. Original manuscript received by the IRE, June 21, 1954. The work herein was performed by the writer while employed by the Sperry Gyroscope Co., and the information is published with their permission.

† Stanford Res. Inst., Stanford, Calif.

[1] G. E. F. Fertel, R. W. L. Batt, J. A. Barrable, and C. S. Wright, British Patent No. 592,224; September 11, 1947.

[2] A. E. Laemmel and H. M. Altschuler, "Final Report of Improved Impedance Measuring Methods Over the Frequency Range of 100–10,000 Mc," MRI Rep.; June 5, 1951. (Air Force Contract W33-038-ac-18115).

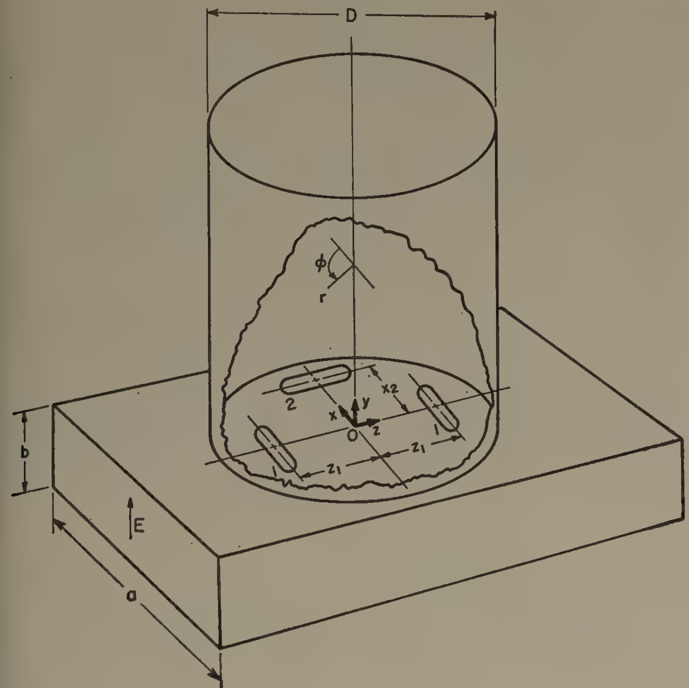

Fig. 1—The circular-polarization coupler.

These two properties make possible a multitude of different methods of measuring impedance. For example, the probe may be operated manually to determine the vswr and minimum position, or it may be rotated at a constant speed and the standing wave displayed on an oscilloscope. With certain further complications, the coupler may also be used to indicate or directly record the impedance as the frequency is swept over the 20-per cent operating range of the instrument. In addition, many other applications requiring phase and amplitude measurement are possible.

A comparison of the new device with slotted-waveguide impedance meters and reflectometers is of interest. As compared to the first, the circular-polarization coupler appears to be less expensive to manufacture, and has the operational convenience of rotary motion with maxima and minima always separated by exactly a quarter of a turn. Because of the rotary motion and constant equivalent wavelength, it is much more easily adapted to automatic instrumentation. As compared to the reflectometer, the circular coupler offers readily available phase data and does not require a well-matched termination. Other advantages will appear in the remainder of this paper. The only important disadvantage appears to be a limitation of the bandwidth to about 20 per cent.

QUALITATIVE EXPLANATION OF THE PRINCIPLE OF OPERATION

The performance of the circular-polarization coupler will first be explained in simple terms, and then the design formulas will be given. In Fig. 1, slots 1 couple the H_x fields in the two waveguides, while slot 2 couples the respective H_z fields. Hence, slots 1 and 2 excite TE$_{11}$-mode waves in the circular waveguide that are orthogonal in space relationship. In addition, assuming a pure traveling wave in the rectangular waveguide, H_x and H_z are 90 degrees out-of-phase, and therefore the two TE$_{11}$ waves are also 90 degrees out-of-phase. These facts, plus an additional stipulation that the amplitudes of the two TE$_{11}$ waves be equal, are well-known conditions for the excitation of a circularly-polarized wave. The equality of the amplitudes may be obtained by properly positioning and dimensioning the slots.

Further thought shows that a TE$_{10}$ wave propagating in the $+z$ direction will produce a right-hand circularly-polarized wave (i.e., clockwise rotation viewed from behind), while a TE_{10} wave propagating in the $-z$ direction will produce a left-hand circularly-polarized wave. If the rectangular guide is connected to a mismatched impedance, waves in both the $+z$ and $-z$ directions are present, and hence both right- and left-hand circularly polarized components exist in the circular wave-guide. The combination of these components is an elliptically polarized wave whose axial ratio is equal to the ratio of the sum and difference of the component amplitudes, which in turn are proportional to the respective TE_{10} amplitudes in the rectangular guide. Therefore, it follows that the axial ratio is equal to the standing-wave ratio in the rectangular waveguide. It is also clear that the angular position of the major or minor axis of the polarization ellipse will vary linearly with the angle of the reflection coefficient of the load, a 360-degree change in the latter producing a 180-degree change in the former. From another viewpoint, the minor axis of the ellipse will vary linearly with the minimum position in the rectangular guide, a one guide-wavelength change in the latter corresponding to a 360-degree change in the former. It can, therefore, be seen that all information concerning vswr, impedance and reflection coefficient of a load connected to the rectangular waveguide may be obtained by means of a probe rotated in a transverse plane of the circular waveguide.

In addition to the arrangement in Fig. 1 many other aperture configurations are capable of producing circular polarization. A few of these possibilities are shown in Fig. 2, page 1556. The first two types are narrow-band designs yielding an axial ratio proportional to λ_g (wave-

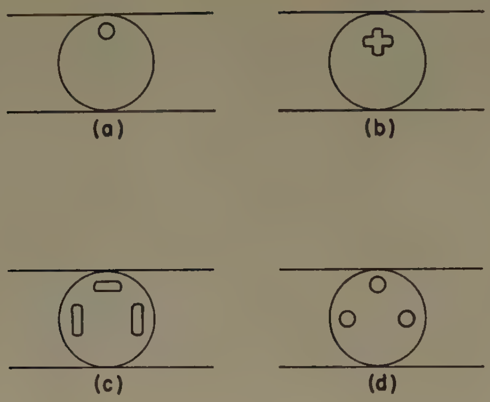

Fig. 2—Aperture configurations for use with the circular-polarization coupler.

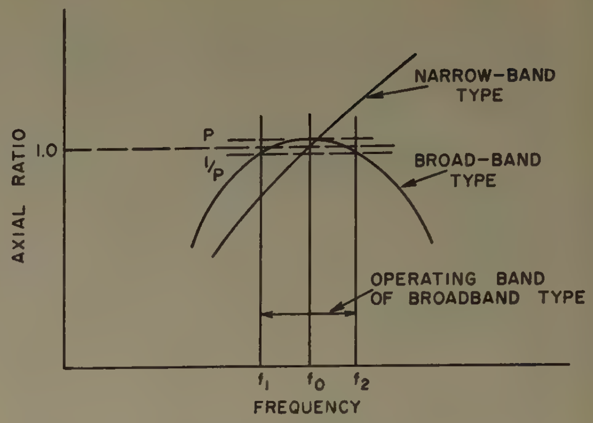

Fig. 3—Axial-ratio vs frequency for the broadband and narrow-band aperture configurations.

length in the rectangular waveguide), as sketched in Fig. 3. The following two types yield a compensated axial-ratio response by means of a proper spacing of apertures along the z axis. In these cases the axial ratio is proportional to

$$\frac{\cos\left(2\pi z_1/\lambda_g\right)}{\lambda_g},$$

which is seen in Fig. 3 to provide a very close approximation to a unity axial ratio over a wide frequency range. Calculation shows that in the case of unity vswr the axial ratio may be held within plus and minus two per cent of unity over approximately a 20-per cent frequency band.

It is characteristic of an elliptically-polarized wave that the major and minor axes are fixed in direction at all transverse planes. The minimum-signal position of the rotatable probe is therefore independent of the particular transverse plane in which the probe is located. However, the probe should be far enough from the apertures to insure that all higher modes excited by the apertures have decayed to negligible amplitudes. A distance of three times the waveguide diameter should be ample for this purpose. By measuring the angular position of the minimum from an axis parallel to the x axis (Fig. 1), the angle of the reflection coefficient at $z=0$ is equal to 2ϕ, and can be read directly (if desired) from a calibrated angle scale concentric with the drive shaft of the probe. This is a convenience unobtainable with a slotted line, due to the variation of guide wavelength with frequency.

A matched load is not necessary at the end of the circular waveguide. It is merely required that the load be symmetrical with respect to the probe. This may be readily understood by representing the elliptical wave by two orthogonal linearly-polarized components, one parallel to the probe and the other at right angles to it. (This can always be done.) If the component parallel to the probe sees a mismatch, repetitive reflections will occur resulting in a certain degree of increase or decrease of the picked-up signal depending on the degree of mismatch and the length of the circular waveguide. Since the end wall containing the small apertures is virtually a perfect short circuit to a wave returning at any angle, the relative change in signal due to the reflections will be independent of the probe angle, and the axial ratio and minimum position indicated by the probe will be unaffected. The component perpendicular to the probe will be reflected repeatedly until it is dissipated. Of course, if the load were not symmetrical with respect to the longitudinal plane through the probe, cross coupling between the components might occur with the result that errors would be introduced. The necessary symmetry is not believed to be difficult to obtain.

Design Relationships

The design formulas for the three-slot circular-polarization coupler have been derived by the writer with the aid of Bethe's small-aperture coupling theory,[3] but the analysis is omitted from this paper in order to avoid excessive length. This method is not an exact one, but it has the advantage of yielding relatively simple equations in closed form.

The coupling through a small aperture is proportional to the "magnetic polarizability" M of the aperture. The ratio of the polarizabilities of the apertures in Fig. 1 yielding circular polarization is given by

$$\frac{M_2}{M_1}=\frac{4x_2aJ_1(\alpha z_1)\cos\dfrac{360z_1}{\lambda_{g0}}}{pz_1\lambda_{g0}J_1(\alpha x_2)\sin\dfrac{180x_2}{a}} \qquad (1)$$

where α is equal to $2\pi/1.705D$, x_2, z_1, a and D are dimensions shown in Fig. 1, $J_1(\alpha z_1)$ and $J_1(\alpha x_2)$ are Bessel functions of the first kind and first order, λ_{g0} is the guide wavelength in the rectangular waveguide at the design frequency, and p is a tolerance factor near unity defined in Fig. 3.

The axial ratio, $A.\ R.$, has the following frequency dependence

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[3] H. A. Bethe, "Lumped Constants for Small Irises," Radiation Lab. Rep. 43-22; March 24, 1943.

$$A.R. \propto \frac{1}{\lambda_g} \cos\left(\frac{360 z_1}{\lambda_g}\right) \tag{2}$$

which is sketched in Fig. 3 for an arbitrary value of p. In the case of $p=1.02$, a tolerance of ± 2 per cent is obtained from $\lambda_g = 8.82 z_1$ to $6.28 z_1$. Hence in this case z_1 should be selected so that the desired guide-wavelength range lies between these limits. The maximum point of the $A.R.$ curve is found by differentiation to occur at $360 z_1/\lambda_{g0} = 49.3$ degrees, or $\lambda_{g0} = 7.30 z_1$. This is the value of λ_{g0} to be used in (1). It occurs very near the center frequency of the compensated band.

The dimension x_2 may be selected arbitrarily, but, of course, must locate slot 2 within the circular waveguide. The diameter D should be chosen so that the operating band lies between the cut-off frequencies of the TE_{11} and TM_{01} modes in the circular waveguide, a 1.307-to-one range. The following equation selects D so that the design frequency lies half way between these cut-off frequencies.

$$D = 0.665 \lambda_0 \tag{3}$$

where λ_0 is the free-space wavelength at the design frequency.

As an example, let $f_0 = 9{,}400$ mc, $p = 1.02$ and $x_2 = 0.320$ inch in a circular-polarization coupler for $1 \times \frac{1}{2}$ inch waveguide.

Then

$\lambda_0 = 1.257$ inches

$D = 0.836$ inch

$a = 0.900$ inch

$\lambda_{g0} = 1.753$ inches

$z_1 = 0.240$ inch

$\alpha = 4.41$ radians per inch

$360 z_1/\lambda_{g0} = 49.3$ degrees

and, with the aid of a table of Bessel functions,

$$\frac{M_2}{M_1} = 1.640.$$

This result stipulates the relative sizes of the apertures. Next, actual sizes must be determined that will provide a satisfactory degree of coupling. Assuming circular polarization, the power coupled through aperture 2 is equal to that through aperture 1, and is equal to the power available to a linearly-polarized pickup probe. The ratio of this power to the rectangular-waveguide power is given by

$$\frac{P_{\text{out}}}{P_{\text{in}}} = \frac{64\pi^2 \lambda_{g0} M_2^2 \sin^2\left(\frac{180 x_2}{a}\right) [J_1(\alpha x_2)]^2}{3 a^3 b D^2 \lambda_{g0}' \alpha^2 x_2^2} \tag{4}$$

where b is the height of the rectangular waveguide and λ_{g0}' is the guide wavelength in the circular waveguide. In the above example,

$$b = 0.400 \text{ inch}$$

$$\lambda_{g0}' = 2.66 \text{ inches}$$

and

$$\frac{P_{\text{out}}}{P_{\text{in}}} = 81.8 M_2^2.$$

Values of the magnetic polarizability of a slot are available.[4] For example, consider a slot with semi-circular ends, the over-all length being denoted by l and the width by w. If $l = 0.300$ inch and $w = 0.094$ inch, then reference 4 gives

$$M_2 = 0.0986 l^3 = 0.00266 \text{ inch}^3$$

and

$$\frac{P_{\text{out}}}{P_{\text{in}}} = 81.8 \times 0.00266^2 = 0.000580$$

which represents a coupling loss of 32.4 db.

In addition, wall thickness and large-aperture effect modify the coupling of an aperture somewhat.[5] If a wall thickness of 0.032 is assumed, the thickness increases the loss by about 5 db, and the large-aperture effect decreases it by about 2 db. The resulting coupling loss should therefore be approximately 35 db.

To complete the above example, $M_1 = M_2/1.640 = 0.00162$. Reference 2 shows that this value of magnetic polarizability may be obtained from a slot having $l = 0.265$ inch and $w = 0.063$ inch.

EXPERIMENTAL TESTS

A circular coupler having the dimensions computed in the above example has been constructed and tested. The circular waveguide contained a contacting rotary joint, and a pickup probe on the wall of the movable portion. [See Fig. 5(a)]. The circular waveguide beyond the probe was terminated by a conical wood plug. With the rectangular waveguide connected to a matched load, the axial ratio was measured by rotating the probe by hand around the circumference of the circular waveguide. The results of the initial test showed a minimum axial ratio of 1.11, indicating that the coupling through the transverse slots was too small by about one db. Also, the center frequency was several per cent lower in frequency than desired. These errors are considered to be reasonable in view of the approximations involved in the derivation of the formulas. After making a few small changes the considerably improved response shown in Fig. 4 (on the following page) was obtained. If desired, a further improvement in performance is believed to be possible.

MEASUREMENT SYSTEMS UTILIZING THE CIRCULAR-POLARIZATION COUPLER

The circular-polarization coupler can be used to measure impedance in a variety of ways ranging from hand operation to fully-automatic Smith-chart recording. Fig. 5(a) shows a simple arrangement suitable for

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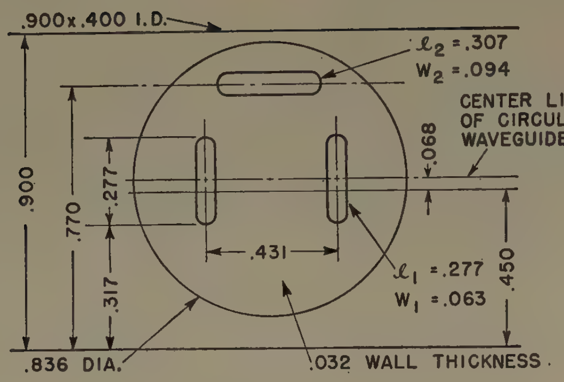
[4] S. B. Cohn, "Determination of aperture parameters by electrolytic-tank measurements," PROC. I.R.E., vol. 39, pp. 1416–1421; November, 1951.

[5] S. B. Cohn, "Microwave coupling by large apertures," PROC. I.R.E., vol. 40, pp. 696–699; June, 1952.


.900x.400 I.D.
ℓ_2 =.307
W_2 =.094
CENTER LINE OF CIRCULAR WAVEGUIDE
.068
.900
.770
.277
.317
.431
ℓ_1 =.277
W_1 =.063
.450
.836 DIA.
.032 WALL THICKNESS

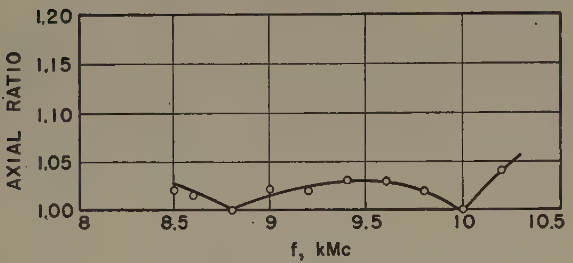


Fig. 4—Response of the latest design.

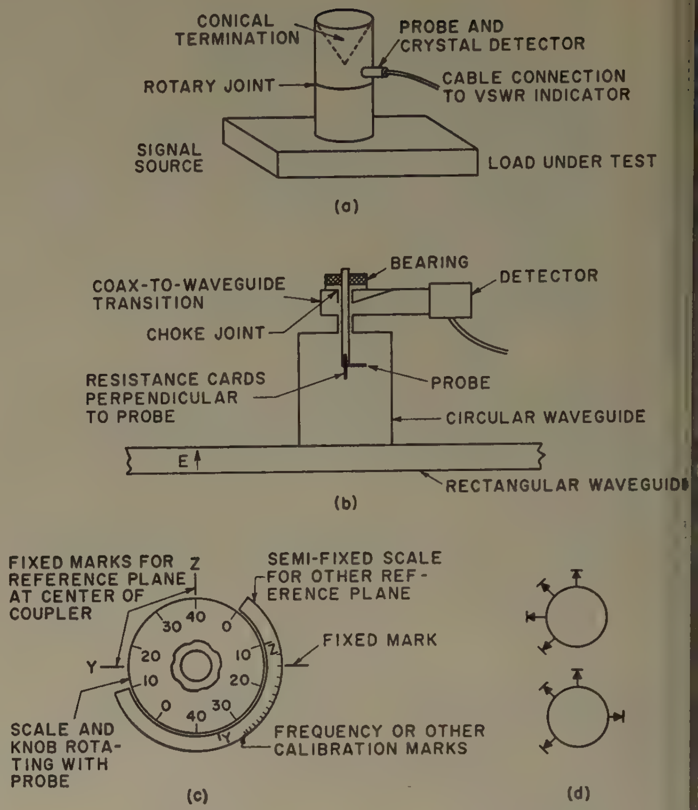

CONICAL TERMINATION
PROBE AND CRYSTAL DETECTOR
ROTARY JOINT
CABLE CONNECTION TO VSWR INDICATOR
SIGNAL SOURCE
LOAD UNDER TEST
(a)
BEARING
DETECTOR
COAX-TO-WAVEGUIDE TRANSITION
CHOKE JOINT
RESISTANCE CARDS PERPENDICULAR TO PROBE
PROBE
CIRCULAR WAVEGUIDE
E
RECTANGULAR WAVEGUIDE
(b)
FIXED MARKS FOR REFERENCE PLANE AT CENTER OF COUPLER
Z
SEMI-FIXED SCALE FOR OTHER REFERENCE PLANE
FIXED MARK
Y
SCALE AND KNOB ROTATING WITH PROBE
FREQUENCY OR OTHER CALIBRATION MARKS
30 40 0 10 20 10 20 0 40 30
(c)
(d)

Fig. 5—(a) and (b) Two circular-coupler designs for impedance measurement; (c) A suitable dial arrangement for hand operation; (d) Probe arrangements for Samuel method.

hand operation. This manner of construction was used successfully in testing the original coupling-aperture designs. The cable connection limits the rotation of the probe, but since two maxima and two minima occur per revolution, the prove need never be rotated more than a half-turn. The probe penetration should be large in order to make up for the 30 db or so attenuation of the apertures. As explained above, the reflection from the probe will not result in any measurement error. The termination may be a conical length of lossy material, or may be a pair of resistance cards intersecting on the axis of the circular guide and mounted parallel and perpendicular to the probe.

In an experimental rotating-probe model [Fig. 5(b)], the probe consisted of silver paint (DuPont No. 4929) on a polystyrene disc. The size of the probe and its spacing from the end wall was selected to give a good match over the frequency range. Lossy resistance cards were mounted on the disc in order to absorb the TE_{11} component at right angles to the probe.

Fig. 5(c) shows a possible dial arrangement for a hand-operated model. The dial is directly calibrated in hundredths of a rotation (or wavelength). Since impedances repeat every half-wavelength, the scale is graduated from 0 to 50 hundredths twice per turn. If the vswr and the dial reading at Z for minimum signal setting is plotted on a Smith chart, the point will indicate the correct impedance at a transverse reference plane in the rectangular waveguide passing through the axis of the circular waveguide. If the dial reading is taken at position Y the plot will show the admittance at the same reference plane. An auxiliary semi-fixed scale may also be provided to give the correct value at another reference plane. This scale must be re-set at each frequency according to a calibration on the scale or on a separate chart.

The circular-polarization coupler can be used to indicate impedance automatically in many ways. For example, it is readily adaptable to A. L. Samuel's four-probe method of automatic Smith-chart indication.[6] Samuel's method has been limited previously to narrow frequency ranges due to the necessary 45-degree spacing of the probes. By placing the probes at 45-degree intervals around the circumference of the circular-polarization coupler [Fig. 5(d)], this disadvantage may be eliminated over a 20-per cent bandwidth, since the 45-degree phase differences will be maintained accurately over the design range of the coupler. Other automatic systems using servo or continuous rotation of a probe are possible, but can not be discussed in this paper.

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[6] A. L. Samuel, "An oscillographic method of presenting impedances on the reflection coefficient plane," Proc. I.R.E., vol. 35, pp. 1279–1283; November, 1947.

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Weighting Functions for Time-Varying Feedback Systems*

J. A. ASELTINE†, ASSOCIATE, IRE, AND R. R. FAVREAU‡, ASSOCIATE, IRE

Summary—The weighting functions for a linear time-varying system provide a means for determining system response to an arbitrary input, and also the mean square output with white noise input. The weighting functions as obtained by direct simulation are not in convenient form for such computation, but can be obtained in such form by use of the adjoint of the original system.

A method is presented for the synthesis of the adjoint system for a given linear time-varying feedback system. The method is based on a physical argument in which outputs due to two inputs which place identical initial conditions in the system are equated.

INTRODUCTION

THE ANALYSIS of linear time-varying systems is inherently more difficult than the analysis of constant coefficient systems. There are two reasons for this. In the first place, the response of a linear time-varying system is a function of two variables, namely, the time at which a disturbance is applied, and the time at which the response is measured. This is contrasted to a constant coefficient system when the response is a function of only one variable, the time difference between the application of a disturbance and the measurement of the response. In the second place, the response of a linear time-varying system must usually be described by more complex functions than the exponential functions which describe the response of constant coefficient systems.

The pencil and paper analysis of linear time-varying systems is usually ruled out because the complex functions which are required to describe their response are either not known or not tabulated. For this reason, and because of time considerations, the analysis of linear time-varying systems is usually performed upon an analog computer. While the computer eliminates the difficulties associated with complicated functions required to describe the response, it does not overcome the added complexity involved, because the response is a function of two variables. In this paper we will describe a technique which may be used to instrument a linear time-varying system upon an analog computer which, to a certain extent, reduces the complexity associated with describing a system response that is a function of two variables.

THE ANALYSIS OF LINEAR SYSTEMS

There are two basic methods of analyzing linear constant coefficient systems. The first method uses the steady-state response to a sine wave of various frequencies to describe the system. The second method, called the weighting-function method, uses the transient response to an impulse or step function to describe the system. Once either of these characteristics is known, the response of the system to any arbitrary input may be determined.

Either of these techniques may be used to analyze linear time-varying systems. However, the transient response method is by far the simplest when an analog computer is used in the analysis. There are two reasons for this. In the first place, the frequency response of a linear time-varying system is a function of time. If the response of the system is to be analyzed at more than one instant, a variety of frequency response curves is necessary. In the second place, an analog computer operates in the time domain, and the obtaining of frequency response characteristics, while possible, represents a long and tedious procedure. In addition, frequency response curves cannot be operated upon by the computer for further system analysis.

The transient response method for linear time-varying systems lends itself quite readily to an analog computer because a time-varying system usually appears to be in the transient state. In addition, the impulse or step function responses which are used to characterize the system are functions of time, and can therefore be operated upon by the computer for further analysis.

WEIGHTING FUNCTIONS

We will call the response of a system to a unit impulse, $\delta(t)$, or any of its derivatives or integrals, a weighting function. For a *fixed* linear system the response, $\theta_0(t)$, to a general input, $\theta_i(t)$ applied at $t=0$, can be written[1]

$$\theta_0(t) = \int_0^t \theta_i(\tau) h_{-1}(t-\tau) d\tau \tag{1}$$

where $h_{-1}(t)$ is the response to a unit impulse. The notation for the weighting functions is tabulated below:

Input	*Response*
doublet $[\delta'(t)]$	h_{-2}
unit impulse $[\delta(t)]$	h_{-1}
step $[u(t)=\int_0^t \delta(t) dt]$	h_0
ramp $[\int_0^t u(t) dt]$	h_1

When the system under study is time-varying, we must specify both the time of application of the input and the time of observation of the response. Thus, we

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† Formerly Res. and Dev. Guided Missile Div., Hughes Aircraft Co., Culver City, Calif.; now, Lear, Inc., Santa Monica, Calif.

‡ Formerly Res. and Dev. Guided Missle Div., Hughes Aircraft Co., Culver City, Calif.; now, Electronics Associates, Inc., Long Branch, N. J.

[1] M. F. Gardner, and J. L. Barnes, "Transients in Linear Systems," John Wiley and Sons, Inc., New York, N. Y.; 1948.

speak of $h_{-1}(t, t_1)$ as the response observed at time t to an impulse applied at time t_1. The counterpart of (1) for a time-varying system (assumed to be at rest for $t<0$) is[2]

$$\theta_0(t) = \int_0^t \theta_i(t_1) h_{-1}(t, t_1) dt_1. \tag{2}$$

If θ_i is white noise with spectral density $W_0(t)$, the standard deviation, σ, of the output may be obtained from the impulse response,[3]

$$\sigma^2(t) = \frac{1}{2}\int_0^t W_0(t_1)[h_{-1}(t, t_1)]^2 dt_1. \tag{3}$$

In practice, θ_i is usually assumed to be a Gaussian variable. Then, since (2) represents a linear operation, θ_0 will be a Gaussian variable[4] and the response can be described in terms of two parameters, mean and standard deviation.

Eqs. (2) and (3) indicate the desirability of having the weighting function, h_{-1}, available. However, if we attempt to use an analog computer to perform the above mathematical operations we run into some difficulty. The reason for this is that the integrating procedures described above are with respect to t_1, the time at which the impulse is applied. If our system is simulated upon the computer and an impulse introduced, the resulting weighting function is a function of t (the time of measurement of the response) for a particular value of t_1. The computer could be operated a number of times, always introducing the impulse at a different time t_1 and measuring the response at a particular value of t. The weighting function, $h_{-1}(t, t_1)$, could then be determined point by point for a particular value of t as a function of t_1. This procedure is long and does not allow the computer to perform the integrations described in (2) and (3), except on a greatly expanded time scale.

The analog of a linear time-varying system, therefore, does *not* provide us with the desired weighting function for easy system analysis. More desirable would be a system which provides us with the impulse response as a function of the time at which the impulse is introduced. This new system is called the *adjoint* system, and its validity and instrumentation have been described by Laning and Battin.[5]

Although the computer would probably not be used to perform the operation in (2), the weighting functions available as a function of time of application are very useful in predicting system behavior. Instrumentation of (3) on the computer, on the other hand, represents a large saving in time over methods that do not use the adjoint system.

We will show how the adjoint system can be synthesized from the original system block diagram, using relations between various weighting functions. Long mathematical argument is avoided so that the physical nature of the problem becomes more apparent. A short appendix contains a discussion of the mathematical aspects of the problem, but this material is not necessary to the development.

The synthesis of the adjoint system is accomplished in the following way:

1. The response of a system to a set of initial conditions is independent of the manner in which these initial conditions are introduced into the system. It will be shown for a time-varying system that an initial condition can be introduced
   (a) by applying appropriate impulse functions at the input
   (b) by applying a constant input at an appropriate place in the system.
2. Since excitations (a) and (b) above produce the same result at the output, these outputs, which can be written in terms of weighting functions, are equated.
3. By using derivative relations (developed in the next section) between weighting functions, the equation resulting can be written as a differential equation with impulse response as dependent variable. The derivatives are with respect to t_1, the time of application; the differential equation is shown to be that for the adjoint system.

PRELIMINARY RESULTS

It is the existence of derivative relations with respect to t_1 among the weighting functions that makes possible a system with t_1 as the independent variable leading to instrumentation of (2) and (3) on the analog computer. The relation between the response to a ramp (h_1) and the response to a step (h_0) is derived as follows.

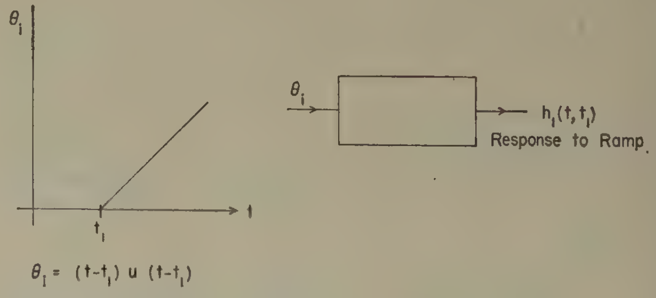

Fig. 1—Response of a linear time-varying system to a ramp.

In Fig. 1, a system is subjected to a ramp at time t_1. The response is $h_1(t, t_1)$. In Fig. 2, θ_i is made up of the difference of two ramps of magnitude $1/\Delta t_1$, the second ramp being applied at time $t_1+\Delta t_1$. Because the system is linear and superposition holds, the response is also a difference. Now, we note that as Δt_1 becomes small,

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[2] L. A. Zadeh, "Frequency analysis of variable networks," PROC. I.R.E., vol. 38, pp. 291–299; March, 1950.

[3] R. R. Bennett, "Analog computing applied to noise studies," PROC. I.R.E., vol. 41, pp. 1509–1513; October, 1953.

[4] $\theta_0(t)$ will not be stationary, but at a given time, t_2, it will be Gaussian over the ensemble if $\theta_i(t_2)$ is Gaussian.

[5] J. H. Laning, Jr. and R. H. Battin, "On an Application of the Use of Analog Computers to Methods of Statistical Analysis," Project Cyclone Symposium II, Reeves Instrument Corp., New York, N. Y.; 1952.

$$\theta_i \to u(t - t_1), \text{ unit step}$$

and

$$\text{response} = \lim_{\Delta t_1 \to 0} \frac{h_1(t, t_1) - h_1(t, t_1 + \Delta t_1)}{\Delta t_1}$$

$$= -\frac{\partial h_1(t, t_1)}{\partial t_1}.$$

Therefore

$$h_0(t, t_1) = -\frac{\partial}{\partial t_1} h_1(t, t_1).$$

The general relation is

$$h_n(t, t_1) = \frac{\partial}{\partial t_1} h_{n+1}(t, t_1). \tag{4}$$

We note that these derivative relations between the weighting functions are all with respect to t_1, the time at which a disturbance is applied. These relationships apply to all linear time-varying systems. It would now seem reasonable to look for some relationship between the weighting functions that is particular to the system to be analyzed. If this can be done we may then formulate a differential equation with the desired impulse response as the dependent variable and the time at which a disturbance is applied, t_1, as the independent variable.

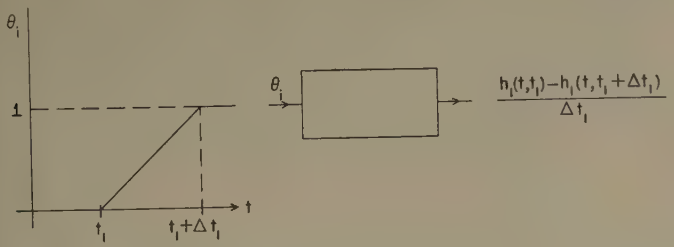
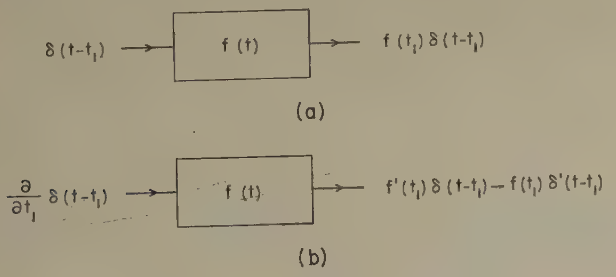


Fig. 2—Response of linear time-varying system to the difference of two ramps approximating a step function.


$\delta(t-t_1)$
f (t)
$f(t_1)\delta(t-t_1)$
$\frac{\partial}{\partial t_1}\delta(t-t_1)$
f (t)
$f'(t_1)\delta(t-t_1)-f(t_1)\delta'(t-t_1)$

(a)

(b)

Fig. 3—Response of time-varying gain element to impulse and doublet functions.

As was indicated in the previous section, part of the synthesis of the adjoint depends on the application of appropriate impulse functions at the input of a time-varying system. The response of a time-varying gain element to an impulse applied at time t_1 is shown in Fig. 3(a). The response of the variable gain element to a doublet is found by differentiating input and output of Fig. 3(a) with respect to t_1 (a procedure justified by the superposition property). The result is shown in Fig. 3(b).

TIME-VARYING FEEDBACK SYSTEM

Consider the feedback system in Fig. 4. Here $1/p$ is the transfer function representing integration. There are two integrations in the above loop and the system in general would have two initial conditions. Let us determine two different driving functions which are equivalent to a unit initial condition at point B at time t_1. We will then equate the responses to these two inputs to form an equation describing the system.

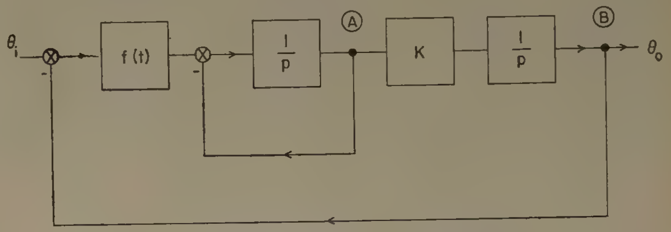


Fig. 4—Linear time-varying feedback system.

Suppose we have a δ function at time t_1 at the input:

$$\theta_i(t) = \delta(t - t_1). \tag{5}$$

Referring to Fig. 4, we see that this becomes $f(t_1)\delta(t-t_1)$ at the output of the $f(t)$ block, which is integrated to become an initial condition of magnitude $f(t_1)$ at point A. The initial value at B will be zero.

Next consider a doublet at time t_1 at the input:

$$\theta_i(t) = \delta'(t - t_1) = -\frac{\partial}{\partial t_1}\delta(t - t_1). \tag{6}$$

The output of the $f(t)$ block is [from Fig. 3(b)]

$$f(t_1)\delta'(t - t_1) - f'(t_1)\delta(t - t_1).$$

The input (6) produces an impulse of magnitude $f(t_1)$ and an initial condition of magnitude $-f'(t_1)$ at point A. The impulse continues through the circuit to produce an initial condition of magnitude $Kf(t_1)$ at point B and magnitude $-f(t_1)$ at point A. The input required to produce a unit initial condition at point B alone can be made up of a combination of (5) and (6) in the following manner. To obtain the desired unit initial condition at B we need, referring to Fig. 5 on the following page, an input of

$$\frac{1}{Kf(t_1)}\delta'(t - t_1).$$

But this places at point A the initial condition:

$$-\frac{f'(t_1)}{Kf(t_1)} - \frac{1}{K} = -\frac{1}{K}\left(1 + \frac{f'(t_1)}{f(t_1)}\right).$$

Now this can be cancelled if we introduce a δ function at the input of magnitude

$$\frac{1}{f(t_1)}\left[\frac{1}{K}\left(1 + \frac{f'(t_1)}{f(t_1)}\right)\right].$$

Therefore, the total input required to place a unit initial

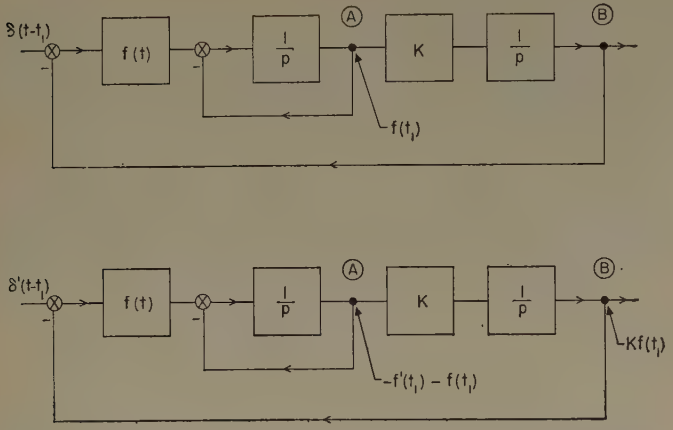


Fig. 5—Initial conditions introduced by impulse and doublet inputs

$$\left[\delta'(t-t_1)=\frac{\partial}{\partial t}\delta(t-t_1)\right].$$

condition at B is

$$\theta_i=\frac{1}{Kf(t_1)}\delta'(t-t_1)+\frac{1}{Kf(t_1)}\left(1+\frac{f'(t_1)}{f(t_1)}\right)\delta(t-t_1). \quad (7)$$

The response due to this input can be expressed in terms of the weighting functions previously defined:

$$\theta_0=\frac{h_{-2}(t,t_1)}{Kf(t_1)}+\frac{h_{-1}(t,t_1)}{Kf(t_1)}+\frac{f'(t_1)}{Kf^2(t_1)}h_{-1}(t,t_1). \quad (8)$$

A *second* and different way to put a unit initial condition at point B is to introduce a unit-step function as shown in step 1 of Fig. 6. This input can be moved around the loop as in steps 2 and 3 of Fig. 6 without changing the output. The net response, as in step 3 is

$$\theta_0=u(t-t_1)-h_0(t,t_1). \quad (9)$$

Now, since this response (9) is due to a unit initial condition at B, as was (8), the two can be equated. We call this the method of equivalent inputs. The resulting equation is

$$\frac{h_{-2}(t,t_1)}{Kf(t_1)}+\frac{h_{-1}(t,t_1)}{Kf(t_1)}+\frac{f'(t_1)}{Kf^2(t_1)}h_{-1}(t,t_1)$$
$$=u(t-t_1)-h_0(t,t_1) \quad (10)$$

or, collecting terms

$$-\frac{1}{K}\frac{\partial}{\partial t_1}\frac{h_{-1}(t,t_1)}{f(t_1)}+\frac{h_{-1}(t,t_1)}{Kf(t_1)}$$
$$+h_0(t,t_1)=u(t-t_1). \quad (11)$$

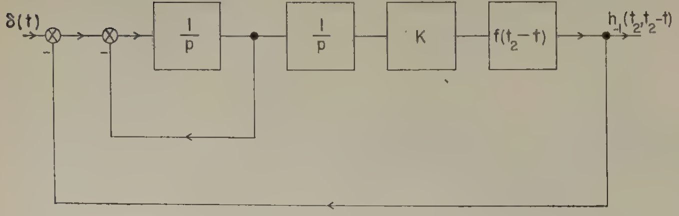


Fig. 7—Block diagram representation of (13).

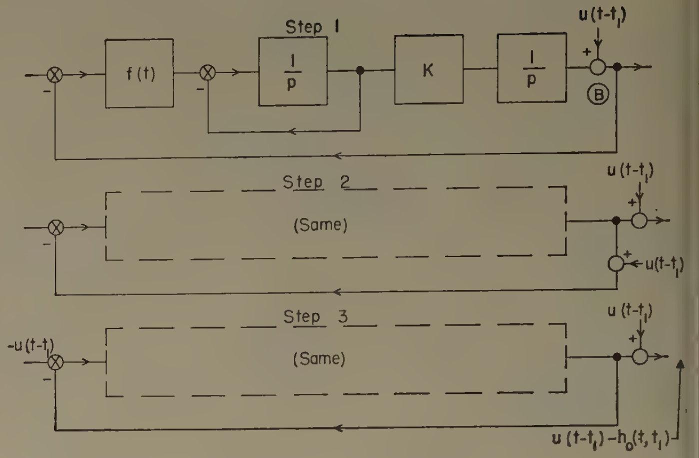


Fig. 6—Steps leading to inputs equivalent to initial condition at point B.

We note in passing that (10) gives a relation among weighting functions that can be used to find a third when two are already known. Now, differentiating with respect to t_1 and letting t_2 be the time of observation, (11) becomes

$$-\frac{1}{K}\frac{\partial^2}{\partial t_1^2}\frac{h_{-1}(t_2,t_1)}{f(t_1)}+\frac{1}{K}\frac{\partial}{\partial t_1}\frac{h_{-1}(t_2,t_1)}{f(t_1)}$$
$$-h_{-1}(t_2,t_1)=-\delta(t_2-t_1). \quad (12)$$

This is a differential equation with the impulse response function as the dependent variable and t_1, the time at which an impulse is introduced, as the independent variable. The time of observation t_2 is a parameter of the equation. In order to put (12) into a form that is suitable for computer instrumentation, we make the change of variable,

$$t=t_2-t_1$$

where t is real computer time. Then (12) becomes

$$\frac{1}{K}\frac{d^2}{dt^2}\frac{h_{-1}(t_2,t_2-t)}{f(t_2-t)}+\frac{1}{K}\frac{d}{dt}\frac{h_{-1}(t_2,t_2-t)}{f(t_2-t)}$$
$$+h_{-1}(t_2,t_2-t)=\delta(t). \quad (13)$$

A block diagram of this equation is shown in Fig. 7, and, rearranging, we have finally the block diagram in Fig. 8.

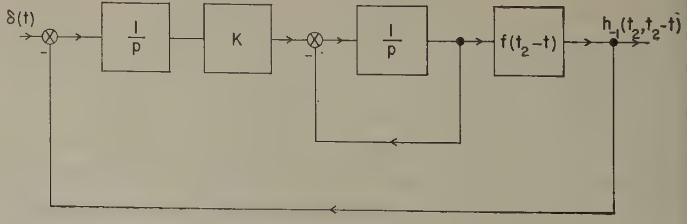


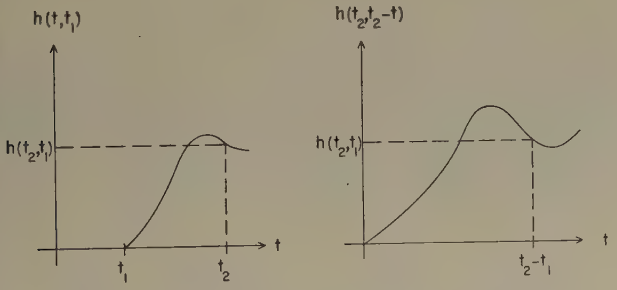
Fig. 8—Adjoint system for Fig. 4.

Comparing Fig. 8 with Fig. 4, we note the following:

1. The order of elements around the loops is reversed.
2. The time varying element in Fig. 8 starts with the value $f(t_2)$ which was its final value in Fig. 4. The argument (t_2-t) decreases as time goes on.
3. The positions of system inputs and outputs are reversed in the two figures. The inputs to both systems are impulse functions.
4. There is available in the system of Fig. 8, the function $h_{-1}(t_2, t_2-t)$. This is one of the weighting functions, but measured at $t=t_2$ for varying-time of application of the impulse. This is a function that can be used in the way described in the introduction to find the response to a general input, or the mean square response to a statistical input.

The properties listed above are those of the adjoint system described by Laning and Battin.[6] The adjoint, therefore, has then been arrived at through a physical argument. A mathematical discussion of the adjoint and an example of the method of drawing the block diagram for an adjoint system will be found in the appendix.

It is interesting to plot representative weighting functions produced by the two systems—the original, Fig. 4, and the adjoint, Fig. 8. These might appear as shown in Fig. 9.



original system adjoint system

Fig. 9—Representative responses of time-varying system and its adjoint.

CONCLUSION

The method discussed has led to a development of the adjoint system by a physical argument, avoiding the discussion of Green's functions and the like. It may, therefore, provide more physical intuition in understanding the adjoint system. Eq. (10) can be used to find other weighting functions graphically from functions already computed.

The system used for illustration of the method can be changed to fit specific problems. However, the adjoint system can be drawn immediately, and the counterpart of (10) found from it.

ACKNOWLEDGMENT

The authors wish to thank R. K. Roney of Hughes Aircraft Company for his part in discussions relating to this work.

APPENDIX

It has been shown[7] that the adjoint system may be used to obtain the weighting function for a linear time-varying system as a function of the time of application of the impulse. In this appendix we will show for a simple system how the adjoint is obtained from the block diagram; how the equation for the adjoint system is related to that for the original system; and how the solutions of the two-systems equations are related.

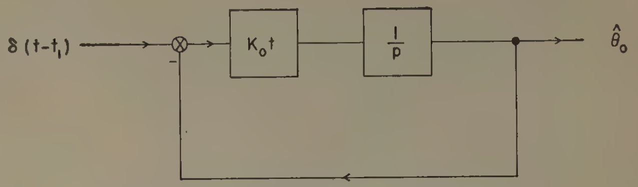


Fig. 10—Simple time-varying feedback system.

Let us consider the servo with varying gain shown in Fig. 10. The input is an impulse function at time t_1. The equation describing this system is

$$\frac{1}{K_0 t}\,\frac{d\hat{\theta}_0}{dt} = \hat{\theta}_0 = \delta(t - t_1). \tag{14}$$

The system could be simulated as shown in Fig. 11.

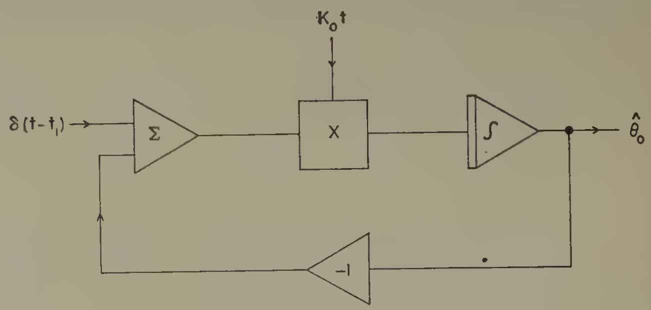


Fig. 11—Analog computer block diagram for Fig. 10.

In Fig. 11, we define the impulse response as

$$\hat{\theta}_0(t) = h_{-1}(t, t_1).$$

Now the adjoint system is found from the original by:

1. Turning each element in the loop around, and reversing the direction of signal flow.
2. Letting the variation of time-varying elements start from some time t_2, and run backwards relative to their action in the original system.
3. Interchanging the input and output of the system. The new input is $\delta(t)$. We will show that the output is $h_{-1}(t_2, t_2-t)$.

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[6] J. H. Laning, Jr. and R. H. Battin, *op. cit.*

[7] J. H. Laning, Jr. and R. H. Battin, *op. cit.*

Following these rules, we have in Fig. 12 the adjoint system for Fig. 11. We will call the impulse response of this new system $\theta_0(t)$. The equation for the system in Fig. 12 is

$$\delta(t) - \theta_0(t) = \frac{d}{dt}\left[\frac{1}{K_0(t_2 - t)}\theta_0(t)\right]. \tag{15}$$

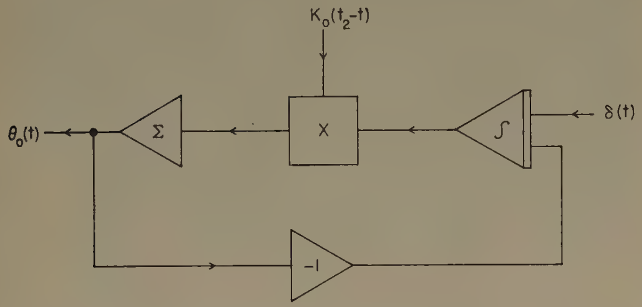

Fig. 12—Adjoint system for Fig. 11.

We change variable, $x = t_2 - t$, $\theta_0(t_2 - x) = \bar{\theta}_0(x, t_2)$, and (15) becomes

$$-\frac{d}{dx}\left[\frac{1}{K_0 x}\bar{\theta}_0(x, t_2)\right] + \bar{\theta}_0(x, t_2) = \delta(x - t_2). \tag{16}$$

This equation is the adjoint[8] of

$$\frac{1}{K_0 x}\frac{d}{dx}h_{-1}(x, t_1) + h_{-1}(x, t_1) = \delta(x - t_1), \tag{17}$$

where we have used h_{-1} for the solution, since the equation is the same as (14). Now a property of the solutions of a pair of adjoint equations like (16) and (17) is (under certain conditions on initial values that are satisfied here)[9]

$$\bar{\theta}_0(\xi_1, \xi_2) = h_{-1}(\xi_2, \xi_1)$$

so that

$$\bar{\theta}_0(x, t_2) = h_{-1}(t_2, x)$$

or

$$\theta_0(t_2 - x) = h_{-1}(t_2, x).$$

Now, letting $x = t_2 - t$, we have

$$\theta_0(t) = h_{-1}(t_2, t_2 - t),$$

which is the desired result. It may be helpful in understanding the relations between h_{-1}, θ_0, and $\bar{\theta}_0$ to compare them with the solutions for the case being considered. The system is described by first order equations which can be solved by quadratures to yield an analytic solution. Solutions to (17), (15), and (16), are

$$h_{-1}(t, t_1) = K_0 t_1 \exp\left(-\frac{K_0}{2}(t^2 - t_1^2)\right)$$

$$\theta_0(t) = K_0(t_2 - t)\exp\left(\frac{K_0}{2}(t^2 - 2t_2 t)\right)$$

$$\bar{\theta}_0(x, t_2) = K_0 x \exp\left(-\frac{K_0}{2}(t_2^2 - x^2)\right).$$

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[8] The adjoint equation is obtained by replacing each operator $f_n(x)(d^n/dx^n)$ by $(-)^n(d^n/dx^n)f_n(x)$.

[9] E. L. Ince, "Ordinary Differential Equations," Dover Publications, New York, N. Y.; 1944. This property and the conditions under which it holds are discussed on pp. 254–256 in connection with the Green's functions for a system and its adjoint. More specifically, the conditions require that the original system and its adjoint *as simulated* are at rest before application of the impulse.

Analysis of Propagating Modes in Dielectric Sheets*

LEONARD HATKIN†, MEMBER, IRE

***Summary*—A procedure is described for calculating the characteristics of propagating modes in dielectric sheets. Specifically, if a required wavelength of the guided wave is given, the appropriate free space wavelength and thickness of the dielectric sheet can be readily obtained, and these variables are related through a single parameter. The methods used are those of geometric optics and transmission line analogies.**

INTRODUCTION

IN RECENT YEARS there has been considerable interest in the propagation of electromagnetic energy in dielectric sheets. Although several solutions for the propagating modes exist in the literature,[1,2] all these solutions suffer from a number of deficiencies. In particular, the most convenient existing solutions involve the simultaneous solution of two equations, one of which is quadratic and the other transcendental.[3] As a result, given a particular geometry and frequency, it is not too difficult to determine the wavelength of the propagating mode. However, it is quite awkward to work in the reverse direction. That is, given a ratio of guide wavelength to free space wavelength, it is not easy to determine the necessary frequency for the geometry involved. The method described below allows this latter objective to be achieved rather readily, and also relates all the variables involved through a single parameter, rather than two parameters, as in the other methods. Although it has been generally recognized that the propagating wave can be broken up into two criss-crossing wave components,[4] this fact has not heretofore been utilized in solving for the propagating modes. The solutions derived here make use of this characteristic, and are obtained by a combination of geometric optics and transmission line analogies.

DISCUSSION

This paper considers only waves of the TE_{no} and TM_{no} types (no variation in the y direction). The geometry involved is shown in Fig. 1. Consider a wave propagating down the dielectric by means of multiple reflections from the dielectric boundaries. This type of wave can be broken up into a wave traveling in the x direction (the propagating wave), and one traveling in the z direction. For propagation to take place, the air region must be below cut-off (angle of incidence must be equal to or greater than the critical angle), thus setting up a standing wave in the z direction. From the geometry and physics of the situation, a number of relationships become apparent:

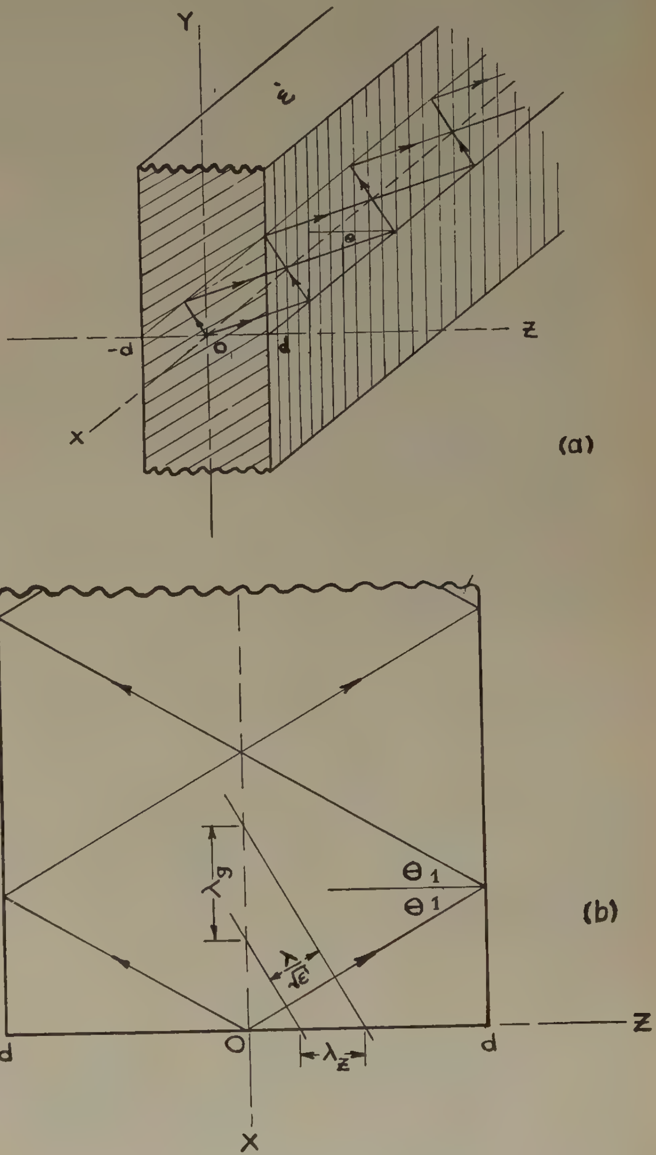

Fig. 1—Dielectric sheet geometry.

$$\lambda/\lambda_x = \lambda/\lambda_g = \sqrt{\epsilon'}\sin\theta_1 \quad (1)$$

$$\lambda/\lambda_z = \sqrt{\epsilon'}\cos\theta_1 \quad (2)$$

$$\sin\theta_2 = \sqrt{\epsilon'}\sin\theta_1 \quad (3)$$

$$Z_{zd}{}^{TE} = E_y/H_x = Z/\sqrt{\epsilon'}\cos\theta_1 \quad (4)$$

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* Decimal classification: R112.33×R118. Original manuscript received by the IRE, April 15, 1954; revised manuscript received, June 23, 1954.

† Evans Signal Laboratory, Belmar, N. J.

[1] J. A. Stratton, "Electromagnetic Theory," McGraw-Hill Book Co., New York, N. Y., p. 590, ex. 9; 1941.

[2] S. A. Schelkunoff, "Electromagnetic Waves," D. Van Nostrand Co., New York, N. Y., pp. 428–431; 1943.

[3] W. O. Schumann, "Elektrische Wellen," Carl Hanser Verlag, Munich, Ger., pp. 187–242; 1948.

[4] S. A. Attwood, "Surface wave propagation over a coated plane conductor," *Jour. Appl. Phys.*, vol. 22, pp. 504–509; April, 1951.

$$Z_{za}^{TE} = Z/\cos\theta_2 = Z/\sqrt{1-\epsilon'\sin^2\theta_1} \quad (5)$$

$$Z_x^{TE} = E_y/H_z = Z/\sqrt{\epsilon'}\sin\theta_1 \quad (6)$$

$$Z_{zd}^{TM} = E_x/H_y = Z\frac{\cos\theta_1}{\sqrt{\epsilon'}} \quad (7)$$

$$Z_{za}^{TM} = Z\cos\theta_2 = Z\sqrt{1-\epsilon'\sin^2\theta_1} \quad (8)$$

$$Z_x^{TM} = E_z/H_y = Z\frac{\sin\theta_1}{\sqrt{\epsilon'}} \quad (9)$$

where

λ is the free space wavelength,
$\lambda_x=\lambda_g$ is the wavelength of the propagation mode
λ_z is the wavelength of the wave traveling in the z direction in the dielectric
ϵ' is the relative dielectric constant of the dielectric medium
θ_1 is the angle of incidence
θ_2 is the angle of refraction (imaginary in the cases here considered)
Z is the intrinsic impedance of free space
Z_{zd}^{TE} is the impedance of the *TE* wave traveling in the z direction in the dielectric
Z_{za}^{TE} is the impedance of the *TE* wave traveling in the z direction in the air region (imaginary in the cases here considered)
Z_x^{TE} is the impedance of the *TE* wave propagating in the x direction
Z_{zd}^{TM} is the impedance of the *TM* wave traveling in the z direction in the dielectric region
Z_{za}^{TM} is the impedance of the *TM* wave traveling in the z direction in the air region (imaginary in the cases here considered)
Z_x^{TM} is the impedance of the *TM* wave propagating in the x direction.

The analysis is similar to that sometimes used for modes propagating between parallel conducting plates. However, whereas for the parallel plate case the angle of incidence could vary from 0 to 90°, in the case of dielectric sheets the angle of incidence must be restricted to angles between the critical angle,

$$\theta_c = \sin^{-1\prime}/\sqrt{\epsilon'}$$

and 90°, so that the waves may be totally reflected and guided by the dielectric sheet. In keeping with the analogy, incidence at the critical angle corresponds to the cut-off frequency and incidence at 90° corresponds to infinite frequency.

The reflection coefficient at the dielectric interface for the transverse field component which is parallel to the interface is

$$R^{TE} = \frac{Z_{za}^{TE}-Z_{zd}^{TE}}{Z_{za}^{TE}+Z_{zd}^{TE}} = \frac{\sqrt{\epsilon'}\cos\theta_1 + j\sqrt{\epsilon'\sin^2\theta_1 - 1}}{\sqrt{\epsilon'}\cos\theta_1 - j\sqrt{\epsilon'\sin^2\theta_1 - 1}} \quad (10)$$

for the E_y component of the *TE* waves, and

$$R^{TM} = \frac{Z_{zd}^{TM}-Z_{za}^{TM}}{Z_{zd}^{TM}+Z_{za}^{TM}}$$

$$= \frac{\cos\theta_1/\sqrt{\epsilon'} + j\sqrt{\epsilon'\sin^2\theta_1 - 1}}{\cos\theta_1/\sqrt{\epsilon'} - j\sqrt{\epsilon'\sin^2\theta_1 - 1}} \quad (11)$$

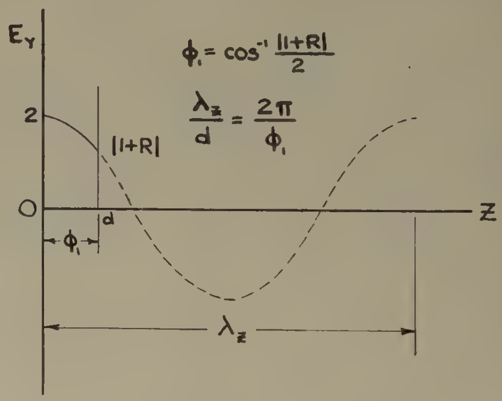
for the H_y component of the *TM* waves. The $\sqrt{-1}$ has been taken as $-j$ in order to cause an exponential decay of the fields in the air region.

These reflection coefficients always have a magnitude of one in keeping with the principle of total reflection, and a phase angle which varies from 0 at $\theta_1=\theta_c$ to π at $\theta_1=90°$. Thus a transverse standing wave will be set up in the dielectric which is co-sinusoidal for the odd numbered modes and sinusoidal for the even numbered modes. The electrical length ϕ_1 of the standing wave from the midplane, $z=0$, to the dielectric boundary, $z=d$, for the lowest mode (TE_{10} or TM_{10}) can be obtained from the expression

$$\phi_1 = \cos^{-1}\left|\frac{1+R}{2}\right|, \quad 0 \leqq \phi_1 \leqq \pi/2. \quad (12)$$

R is determined from (10) or (11), depending on whether we have a *TE* or *TM* wave. The electrical length of the standing wave for the nth order mode (TE_{no} or TM_{no}) can be found by adding appropriate numbers of quarter wavelengths. Thus

$$\phi_n = \phi_1 + (n-1)\pi/2 \quad n = 1, 2, 3, 4, \text{etc.} \quad (13)$$



E_Y
$\phi_1 = \cos^{-1}\frac{|1+R|}{2}$
$\frac{\lambda_z}{d} = \frac{2\pi}{\phi_1}$
2
|1+R|
O
d
Z
ϕ_1
λ_z

Fig. 2—Determination of λ_z from R.

From Fig. 2, the relationship

$$\lambda_{zn}/d = 2\pi/\phi_n \quad (14)$$

is also apparent, where λ_{zn} is the wavelength of the standing wave in the z direction for the nth mode in the dielectric, and 2d is the thickness of the dielectric. Combining (2) and (14) we have

$$\lambda/2d = \frac{\pi\sqrt{\epsilon}\cos\theta_1}{\phi_n}. \quad (15)$$

At cut-off $\phi_1=0$, and $\cos\theta_1=\sqrt{(\epsilon_1'-1)/\epsilon'}$, so that

$$\lambda_{cn}/2d = \frac{2}{n-1}\sqrt{\epsilon'-1}, \quad n = 1, 2, 3, 4, \text{etc.} \quad (16)$$

where λ_{cn} is the cut-off wavelength for the nth mode.

Thus, given a $\lambda/\lambda g$, the appropriate $\lambda/2d$ may be readily found through the single parameter θ_1. Families

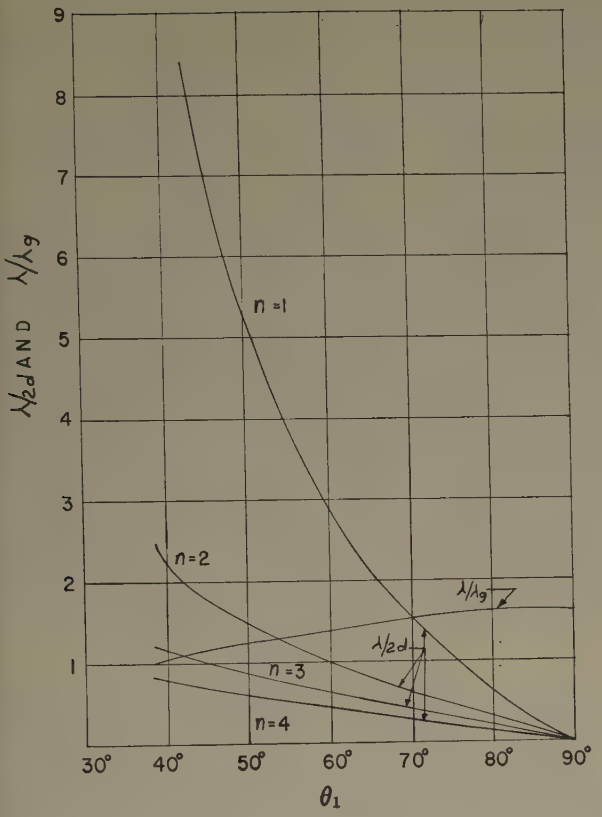


Fig. 3—Plot of $\lambda/\lambda g$ and $\lambda/2d$ vs. θ_1 for *TE* waves in Lucite sheet.

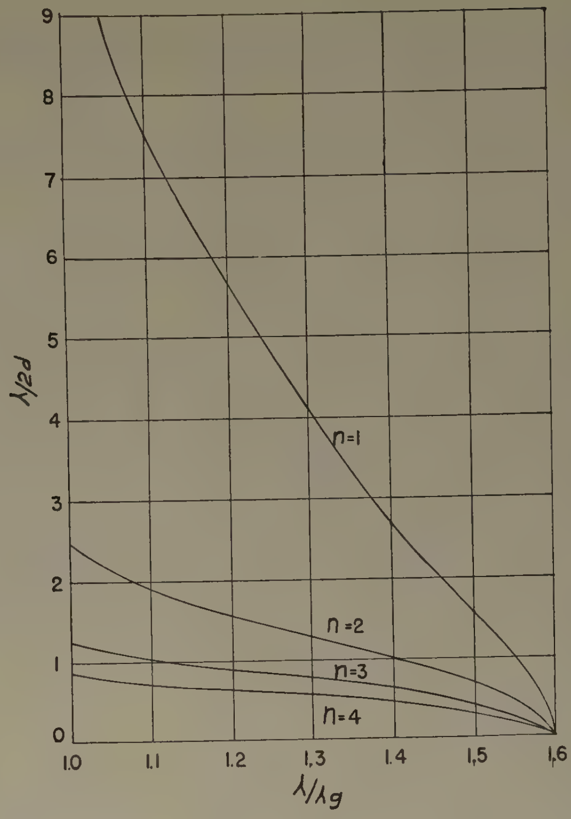


Fig. 4—Plot of $\lambda/2d$ vs. $\lambda/\lambda g$ for *TE* waves in Lucite sheet.

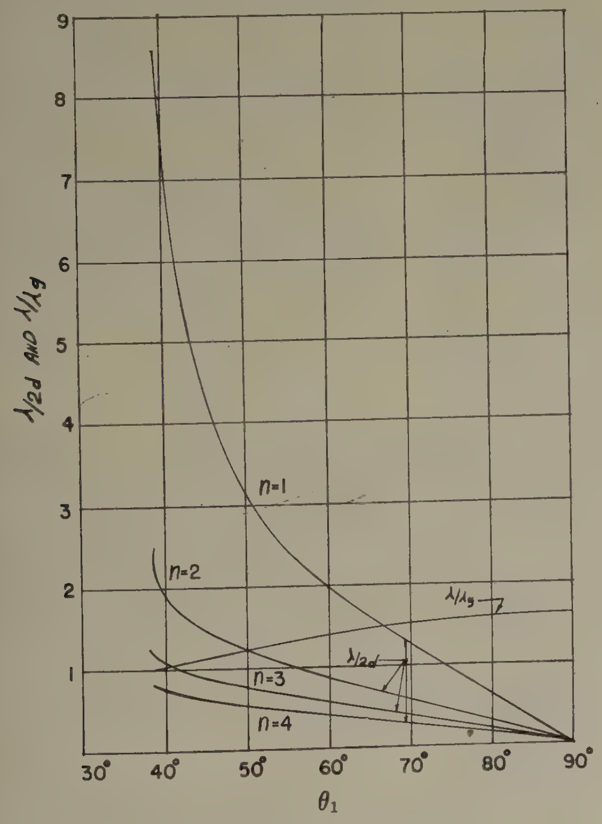


Fig. 5—Plot of $\lambda/\lambda g$ and $\lambda/2d$ vs. θ_1 for *TM* waves in Lucite sheet.

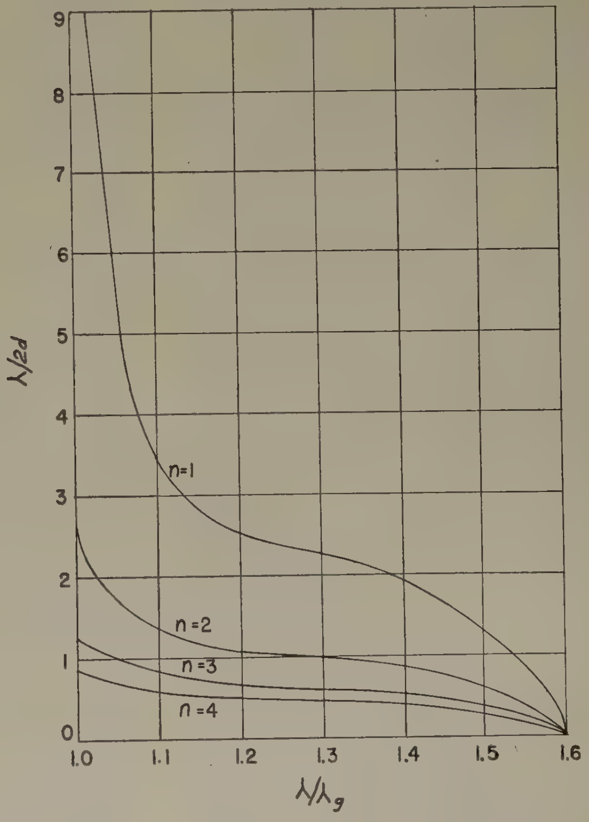


Fig. 6—Plot of $\lambda/2d$ vs. $\lambda/\lambda g$ for *TM* waves in Lucite sheet.

of curves relating the variables $\lambda/\lambda g$ and $\lambda/2d$ may be drawn as in Fig. 3 and 4, which show the *TE* wave curves, and Figs. 5 and 6, which show the *TM* wave curves. These data have been calculated for Lucite sheets, assuming $\epsilon^1 = 2.56$. In general, the procedure is to determine θ_1 from the given $\lambda/\lambda g$. Once θ_1 is known, ϕ_n may be readily found and then $\lambda/2d$ calculated. It should be noted that although the cut-off relations are the same for both the TE and TM waves, elsewhere the curves are different because the effect of ϵ^1 is different in the two cases.

The propagation constant for the wave traveling in the z direction in the air region is imaginary and is given by

$$\frac{2\pi}{\lambda_{za}} = \frac{2\pi}{\lambda}\cos\theta_2 = -j\,\frac{2\pi}{\lambda}\sqrt{\epsilon'\sin^2\theta_1 - 1}\,\cdot \qquad (17)$$

The $-j$ gives rise to the exponential decay characteristic of evanescent waves. The field equations may be obtained by means of transmission line analogies or other standard methods and will not be included here.

With the method described above, analytical relationships between $\lambda/\lambda g$ and $\lambda/2d$ have been rather directly and simply obtained for waves of the TE_{no} and TM_{no} types through the use of a single parameter θ_1, which is the angle of incidence of the criss-crossing traveling wave components at the dielectric to air interface. The salient expressions can be summarized as follows:

$$\lambda/\lambda_g = \sqrt{\epsilon'}\sin\theta_1 \qquad (1)$$

$$\phi_n = \cos^{-1}\left|\frac{1+R}{2}\right| + (n-1)\pi/2 \qquad (12)\text{ and }(13)$$

$$\lambda/2d = \frac{\pi\sqrt{\epsilon'}\cos\theta_1}{\phi_n} \qquad (15$$

For TE waves, R is given by (10) and for TM waves by (11).

CONCLUSIONS

Very often the important parameter in the design o a guided wave system is $\lambda/\lambda g$. The method employe herein allows a direct, analytic determination of th required wavelength and geometry; i.e., $\lambda/2d$. Althoug in this paper the method has been used only to deriv the propagating modes in dielectric sheets, the metho can be extended very simply to calculate the modes i waveguides and transmission lines partially filled wit dielectric—cases which are more conventionally treate by the transverse resonance method.[5] Also, by resorting to geometries other than Cartesian, the cases of dielectric clad conductors, or dielectric tubes of circular o other shape may be solved by this method.

ACKNOWLEDGMENT

The author wishes to acknowledge the helpful discussions with Dr. A. A. Oliner concerning this subject

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[5] D. J. Angelakos, "A Coaxial Line Filled with Two Non-concentric Dielectrics," University of California, Institute of Engineering Research Report, series no. 60, issue no 102; Nov. 9, 1953.

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CORRECTION

Sol Sherr, author of the paper, "Generalized Equations for RC Phase-Shift Oscillators," which appeared on pages 1169–1172 of the July, 1954 issue of the PROCEEDINGS OF THE I.R.E., has brought the following correction to the attention of the editors:

In the drawing for Fig. 5(b), Z_1 becomes Z_1', and Z_1' correspondingly becomes Z_1.

Correspondence

Comment on "A Note on the Analysis of Vacuum Tube and Transistor Circuits,"* L. A. Zadeh[1]

In concluding the above titled paper, Dr. Zadeh remarks that "the method (of constructing admittance or impedance coefficients tables) can readily be extended to networks containing active elements other than three-terminal vacuum tubes and transistors." It seems that although the extension is straightforward as far as admittances (for node equations) are concerned, there arise some essential difficulties in constructing impedance tables (for mesh equations) of elements having more than three terminals.

Consider a general n-terminal network element, and suppose that one of its terminals is used as a ground node in nodal analysis. A total of $(n-1)^2$ admittances are then needed for complete specification of the element. If none of its nodes is grounded, n^2 admittances appear in the analysis, but the added $2n-1$ elements are redundant. because in the complete nxn array the sum of each row and each column is zero.[2] When any other terminal is grounded, the same array of admittances is used, after omitting the appropriate row and column. An admittance table of n^2 elements, with one redundant row and one redundant column, thus contains all the values that may appear in nodal analysis.

In mesh analysis, each mesh concerns a *pair* of terminals, and although not more than n meshes will be needed in any analysis, these n meshes may be chosen out of a set of $\frac{1}{2}n(n-1)$ possible terminal pairs. If the impedance table is to contain all the impedances that may appear in an analysis, it must be constructed in the following manner.

Let I_{ij} be a mesh current entering by terminal i and leaving by terminal j, and let V_{pq} be the voltage terminal between p and terminal q. Then the impedance elements appearing in the mesh analysis are *four-indexed* quantities, as in

$$V_{pq} = Z_{pq,ij}I_{ij},$$

and the impedance "table" should be a four-dimensional array of n^4 elements. The case is not as bad as all that, for both I_{ij} and V_{pq} are skew-symmetrical, so that each of them represents only $\frac{1}{2}n(n-1)$ independent quantities (disregarding the sign) instead of the full n^2, and the impedances may therefore be arranged in a two-dimensional table of $\frac{1}{4}n^2(n-1)^2$ elements. This table is already somewhat incomplete, because when elements are picked out of it to be inserted in mesh equations, the sign of each term is not given in the table; but even thus the table is usually bigger—and more redundant—than the admittance table.

The special case $n=3$ treated in the paper has the simplifying property that

$$n = \tfrac{1}{2}n(n-1)$$

therefore both tables have the same number of elements. (This is analogous to the possibility of representing a skew-symmetrical second order tensor by a vector—but in three dimensions only.) Moreover, in any grounding position, each Z element is equal to one of the Y elements divided by the determinant of the four Y's; and as this determinant is independent of the terminal chosen as ground,[3] all the Z elements are proportional to the Y elements. (Compare Tables II and IV in Zadeh's paper.) Therefore, *in this special case only*, the structure of the Z table is the same as that of the Y table, the sum of each row and column being zero.

These remarks, as well as the fact that in mesh analysis external resistors must sometimes be used, point out that there is not a complete equivalence (or duality) between the methods of mesh and node analysis; the balance being in favor of the latter, at least for the analysis of general n-terminal net work elements.

J. Shekel
8 Ben Yehuda St.
Haifa, Israel

Rebuttal[4]

Mr. Shekel is quite right in stating that there is not a complete equivalence (or duality) between the mesh and node variants of the method under discussion. He is not correct, however, in asserting that the equivalence (or duality) holds only for $n=3$, that is, in the case of three-terminal elements. It holds, more generally, when the network N is planar, and the meshes are chosen in the conventional manner. The number of terminals, n, is immaterial.

As to the fact that in mesh analysis external resistors must sometimes be used (e.g., when dealing with vacuum tubes having infinite grid impedance), this is not really a difference between the two methods, since a similar situation is encountered in the node analysis when one deals with idealized transistors having infinite emitter admittance.

The method described in the note under discussion and in Mr. Shekel's papers has long been used by the writer in his lectures on network theory in Columbia University. In the writer's experience, the use of tables of admittance and impedance coefficients greatly simplifies the analysis of active and passive circuits containing multi-terminal elements. While it is generally true that even in the case of planar networks the analysis is somewhat more readily carried out on the node basis, this fact has no relation to the number of terminals of network elements, as is implied by the discusser.

In conclusion, the writer wishes to thank Mr. Shekel for his interesting comment on the writer's statement regarding the mesh analysis of networks containing n-terminal ($n>3$) elements.

L. A. Zadeh
Dept. of Elec. Eng.
Columbia University
New York 27, N. Y.

Finite Groups in Shannon Coding*

Shannon's well-known formula $C=W\log(1+P/N)$ for the maximum rate of communicating digital information with negligible risk of error is dependent on *sufficiently complicated encoding*.[1] The mathematical derivation in terms of a multidimensional geometric model requires as a necessary condition (though not in itself sufficient) that the number of dimensions in the model-space should tend to infinity, which corresponds to the number of digits in the message-group tending to infinity. Now the delay occasioned by the use of a large number of digits in the message-group may not be prohibitive, but the complexity of the decoding equipment increases with the number of digits in the message-group which must be stored so that the complete message can be identified by correlation with the vocabulary of possible messages. For this reason coding in very large groups may not be practical, and one therefore considers what error rate is to be anticipated with a finite number of digits in the message-group, and how this may be improved by lowering the communication rate below the theoretical maximum.

The crux of the matter is the decoding process, which in Shannon's geometrical model may be described colloquially as follows. We first take our stand on the surface of an x-dimensional hypersphere of radius $\sqrt{[x(P+N)]}$, because with large groups almost all messages will have the maximum mean power (almost all the content of the message hypersphere in many dimensions is at the surface), and the mean noise power N of the system is assumed to be steady and known. We then fish into the interior of the hypersphere with a fishing line of length $\sqrt{(xN)}$ and if the coding at the transmitting end has been properly arranged, there will be only the one correct message within the range of our line.

The requirement that the mean message power shall always have the same value can always be met in the construction of the code: a known example is the error-correcting 7-unit code. With small groups such a code has considerable redundancy (35 characters are provided by a 7-unit code constrained to include always 3 mark units, compared with 128 for an unconstrained code of 7 binary digits), but with extremely large groups the proportion of possible message-groups excluded by the power criterion is small, and the reduction of digit rate by this constraint is readily calculable.

Assuming, then, that we have message power P and white noise with a mean power N, what is the probability that the vector in the Shannon model which represents the combination of noise and message will have the length $\sqrt{[x(P+N)]}$? A qualitative idea of the answer can be obtained by considering separately first the angle between the noise vector and the message vector and then the amplitude of the noise vector.

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* Received by the IRE, February 8, 1954.

[1] Proc. I.R.E., vol. 41, pp. 989–992; August, 1953.

[2] J. Shekel, "Voltage reference node—its transformations in nodal analysis," *Wireless Eng.*, vol. 31, pp. 6–10; January, 1954.

[3] J. Shekel, "Two network theorems concerning change of voltage reference terminal," to be published in Proc. I.R.E.

[4] Received by the IRE, March 3, 1954.

* Received by the IRE, July 29, 1954.

[1] C. E. Shannon, "Communication in the presence of noise," Proc. I.R.E., vol. 37, pp. 10–21; January, 1949.

Thus in Fig. 1 we have assumed two vectors a and b, each of constant length and in ratio 10:1 (i.e., corresponding to signal/noise ratio of 20 db), but with random angle between them. The length of the resultant is denoted by R, and the curves show for various values of x the distribution function $P(R)$, such that the probability of the resultant lying between R and $R+\delta R$ is $P(R)\delta R$. It can be seen that in order to confine the probable resultant within bounds substantially closer than the arithmetic sum and difference of the two vectors, x must be of the order of 100 or more. (For $x=82$; the curve shows that the range of considerable probability is about a quarte of that for $x=2$.) Fig. 2 shows the distribution functions $P(r)$ for the normalized amplitude of the resultant noise vector for several values of x. ($r=1$ is the mos probable value in each case.) Again one ha to go to large values of x before the bound of the range of appreciable probability ar less than plus or minus the rms noise value This adds further uncertainty to the tru magnitude of R when variable noise vector i substituted for constant vector b in Fig. 1.

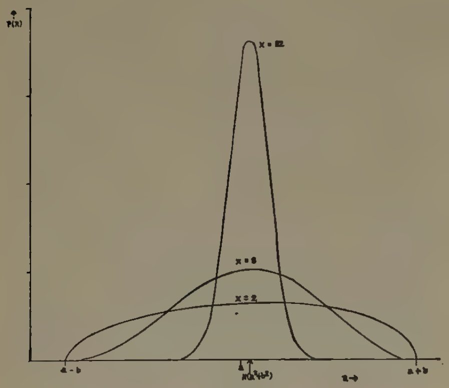

Fig. 1

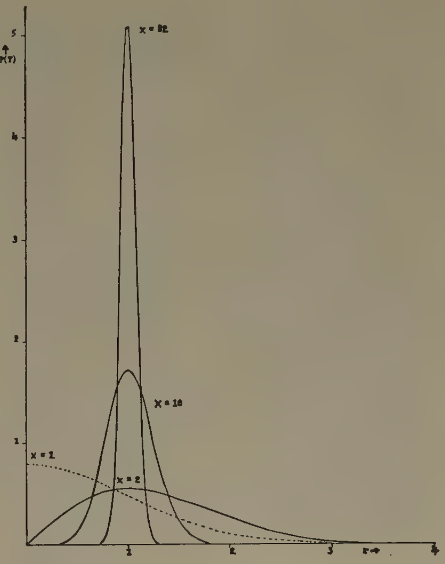

Fig. 2

It appears likely that this need for larg groups is the implicit reason for the neglec of systems based on linear transformation and the philosophy behind Shannon's derivation of the maximum communicatio rate, and the practical development of nonlinear systems such as FM and pulsed systems which employ amplitude limiters an slicers. With white noise the latter are theoretically inferior to an ideal linear system but they are very much simpler to construct

D. A. BELL AND T. C. DUGGAN
Elec. Eng. Dept.
Univ. of Birmingham, Eng.

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[2] Mr. Duggan has a research grant from the Dept of Scientific and Industrial Research.

* * *

Contributors

J. A. ASELTINE

J. A. Aseltine (S'50–A'52) was born on April 12, 1925, in Palo Alto, Calif. He received the B.A. degree in physics in 1947, and the M.S. and Ph.D. degrees in engineering in 1949 and 1952 respectively, from the University of California.

During the period 1943–46 he served with the U. S. Navy, and during 1947–48 was associated with the Radio Laboratory at Consolidated-Vultee in San Diego, Calif.

Now with Lear, Inc., Santa Monica, California, Dr. Aseltine was, from 1952 to 1954, a member of the technical staff at Hughes Aircraft Company, Culver City, Calif. He is a lecturer in engineering at the University of California, Los Angeles.

He is one of the West Coast Representatives of the Professional Group on Circuit Theory, and was 1953–54 Chairman of the Los Angeles Chapter of that Group. He is a member of Pi Mu Epsilon, and Sigma Xi.

H. ATAKA

H. Ataka was born in Kanasawa, Japan, in 1904. He graduated from Tokyo University in 1927, and received the Doctor of Engineering degree in 1936. His doctoral dissertation "On Super-regeneration of an Ultra-Short-Wave Receiver," was published in PROCEEDINGS OF THE I.R.E., pp. 841–884, August, 1935.

During the war, Dr. Ataka was a professor at the Keijyo (Seoul) University, and was advisor to the army on radar, in Korea.

He now holds a professorship at the Ibaraki University, Tagamati, Ibarakiken, Japan.

❖

R. N. BRACEWELL

R. N. Bracewell was born in Sydney, Australia, in 1921. He graduated from the University of Sydney in 1941 with the B.Sc. degree in mathematics and physics, later receiving the degrees of B.E. (1943), and M.E. (1948) with first-class honors. During the war he designed and developed microwave radar equipment in the Radiophysics Laboratory, Sydney. From 1946 to 1949 he was engaged in ionospheric research in the Cavendish Laboratory, Cambridge, England, and there received his Ph.D. degree.

He will be remembered by many American radio scientists as organizing secretary for the 1952 U.R.S.I. Assembly in Sydney. He has since collaborated with Dr. J. L. Pawsey, F.R.S., in writing a book on radio astronomy and now is associated with the Radiophysics Laboratory, Commonwealth Scientific and Industrial Research Organization, Sydney, Australia.

Dr. Bracewell is a Fellow of the Institute of Physics, an Associate Member of the Institution of Electrical Engineers, and a Fellow of the Royal Astronomical Society.

S. B. Cohn (S'41–A44–M'46–SM'51) was born in Stamford, Conn., on October 21, 920. He received the B.E. degree in electrical engineering from Yale University in 1942; the M.S. degree in communication engineering in 1946, and the Ph.D. degree in engineering sciences and applied physics in 1948, both from Harvard University.

S. B. Cohn

From 1942 to 1945 he was employed as a special research associate by the Radio Research Laboratory of Harvard University, also representing that Laboratory as a technical observer with the U. S. Army Air Force in the Mediterranean Theater of Operations. From 1948 to 1953, he was employed by the Sperry Gyroscope Company, where he held the position of research engineer in the Microwave Instruments and Components Department. Since February, 1953, he has been with the Stanford Research Institute, Stanford, Calif. as Head of the Microwave Group of the Radio Systems Laboratory.

He is a member of Tau Beta Pi and Sigma Xi.

❖

R. R. Favreau (S'48) was born in Lowell, Mass., in 1925. He received the B.S. degree in physics from the Massachusetts Institute of Technology, Cambridge, Mass., in 1945.

R. R. Favreau

He then served as a naval officer until 1946, and later joined Raytheon Manufacturing Company. From 1947 to 1954, Mr. Favreau was a member of the technical staff of the Hughes Research and Development Laboratories, Culver City, Calif., where he was in charge of the analog-computer facilities of the Guided Missile Division. Recently he joined Electronic Associates, in Long Branch, N. J.

He is a member of Sigma Xi.

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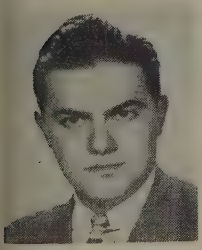
W. L. Firestone (M'49–SM'53) was born in Chicago, Illinois on June 20, 1921. Before World War II he worked at Motorola Inc. as a laboratory assistant and attended night school at the R. C. A. Institute. During World War II he graduated from the Navy Radio Materiel school at Treasure Island, California where he also taught Navy Radio.



W. L. Firestone

He attended the University of California, and later the University of Colorado, receiving his B.S. degree in electrical engineering in June, 1946.

After working on the Manhattan Project for six months he returned to Motorola Inc. as a microwave engineer. He received his M.S.E.E. degree from the Illinois Institute of Technology in June, 1949 and his Ph.D. degree in June, 1952 from Northwestern University. Since 1951 Dr. Firestone has been engaged in vhf research and is Section Head of the Advanced Investigation Section of the Research Department at Motorola Inc.

Dr. Firestone is a member of Sigma Xi, Tau Beta Pi, HKN, Pi Mu Epsilon and A.I.E.E.

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For a photograph and biography of E. L. Ginzton, see page 489 of the February, 1954 issue of the Proceedings of the I.R.E.

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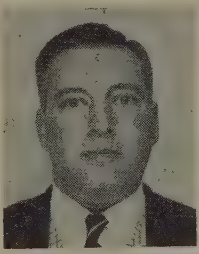
Leonard Hatkin (A'51) was born in New York, N. Y. on March 22, 1920. He received the B.S. degree in Physics *cum laude* from the City College of New York in 1942 and the M.S. degree from Rutgers University in 1950.



Leonard Hatkin

With the exception of a brief tour of active duty with the New Developments Division of the War Department Special Staff during World War II, Mr. Hatkin has been with the Signal Corps Engineering Laboratories continuously since 1942. He is currently Assistant Chief of the Antenna and Microwave Circuitry Section of Evans Signal Laboratory, where he is engaged in the design and development of vhf, uhf, and microwave antennas and their associated transmission line components.

Mr. Hatkin is a member of Phi Beta Kappa and Sigma Xi.

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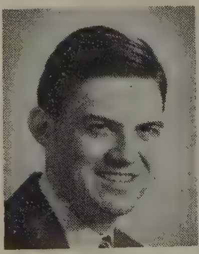
H. O. Hook (S'48–A'51) was born in Atlanta, Georgia on October 5, 1924. He received the B.A. degree with chemistry major from Elon College in 1947, and the B.E.E. degree from North Carolina State College in 1949. Later, in 1950, he received the M.S.E.E. degree from the same College. Since 1950 Mr. Hook has been with the Radio Corporation of America, RCA Laboratories Division.



H. O. Hook

In addition to his activities as a member of I.R.E., he is a member of Sigma Xi.

M. Knoll (SM'51) was born in Schlangenbad, Germany on July 17, 1897. He received ceived the M.S. and Ph.D. degrees in electrical engineering in 1922 and 1924 respectively, from the Institute of Technology in Munich, Germany. From 1927 to 1945 he taught vacuum tube design at the Institute of Technology, Berlin-Charlottenburg, as Assistant, Associate and full Professor, and as Director of its Vacuum Tube Laboratory. After the war, this Laboratory was affiliated with the University of Munich, where as full Professor he taught electron optics and vacuum tube design until 1947.

M. Knoll

From 1932 to 1947 Dr. Knoll was in charge of the Electron Research Laboratory of the Telefunken Corporation, mainly concerned with the development of the television camera, viewing and storage tubes. In 1947 he joined the Evans Signal Laboratories, Belmar, N. J., where he was a consultant for one year. He joined the RCA Laboratories at Princeton, N. J., in December 1948, where he is now engaged in electron research. He is also a lecturer with the rank of full Professor in the Department of Electrical Engineering at Princeton University.

He is a member of Sigma Xi.

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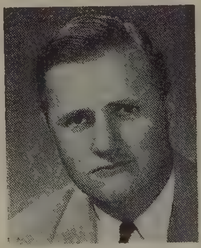
J. M. Lafferty (M'46–SM'48–F'54) was born at Battle Creek, Mich., in 1916. He attended Western Michigan College, and graduated from the University of Michigan in 1939 with a B.S. degree in engineering physics, followed by a M.S. degree in physics in 1940, and a Ph.D. degree in electrical engineering in 1946.



J. M. Lafferty

After aiding in the development of the VT proximity fuse at the Carnegie Institution in Washington, D. C., in 1941, he joined the General Electric Company in 1942, where he has engaged in research on microwave tubes, electron accelerator guns, boride cathodes, and other vacuum-tube problems. Dr. Lafferty now is a research associate in the General Electric Research Laboratory at The Knolls, Schenectady, N. Y.

Dr. Lafferty is a member of Sigma Xi, and received a citation for distinguished alumnus from the University of Michigan in 1953.

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J. R. Macdonald (S'44–A'48–SM'54) was born in Savannah, Georgia on February 27, 1923. He received a B.A. degree in physics from Williams College and an S.B. degree in electrical engineering from M.I.T. in 1944.

J. R. Macdonald

After a semester of teaching at the M.I.T. Army-Navy Technical Radar School, he was commissioned an ensign in the U. S. Naval Reserve in 1944. Upon completing radar courses at the Harvard and M.I.T. radar schools, he served as a technical radar officer. After release to inactive duty in 1946, he returned to M.I.T. and received the S.M. degree in electrical engineering in 1947. During this period he carried out research on storage tubes for the M.I.T. digital computer project. After a year's further graduate study in the M.I.T. Physics Department, he was awarded a Rhodes Scholarship from Massachusetts for study at Oxford University. Dr. Macdonald received a D.Phil. degree in physics from Oxford in 1950 for theoretical and experimental work on ferromagnetic phenomena. In 1950, Dr. Macdonald joined the Physics Department of the Armour Research Foundation and there carried out and directed work in theoretical and experimental physics until 1952. He then spent a years' leave of absence at the Argonne National Laboratory of the A.E.C. working on solid-state physics problems. Dr. Macdonald is presently head of the Physics Research Section of Texas Instruments Incorporated, Dallas, Texas, and is primarily concerned with solid-state physics research.

Dr. Macdonald is a member of Phi Beta Kappa, Sigma Xi, and the American Physical Society.

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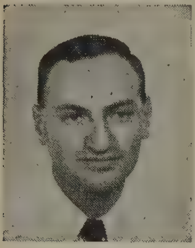

E. J. Nalos

E. J. Nalos was born in Prague, Czechoslovakia, September 10, 1924. He graduated from King Edward High School, Vancouver, Canada, in 1941. In 1946 he received the degree of Bachelor of Applied Science in electrical engineering, and in 1947 the degree of Master of Applied Science from the University of British Columbia. He continued graduate study as a research and teaching assistant at the Microwave Laboratory at Stanford University under a Sperry Gyroscope Company fellowship, obtaining his Ph.D. in 1951. Dr. Nalos is now a Research Associate at the Microwave Laboratory at Stanford University, Stanford, Calif., where he is active in high-power microwave tube research and the student laboratory instruction program.

He is a member of Sigma Xi.

J. L. Stewart

J. L. Stewart (S'48–A'50–M'53) was born in Pasadena, Calif., on April 19, 1925. After serving as an Aerial Navigator in the U. S. 8th Air Force during World War II, he returned to school where he received the B.S., M.S., and Ph.D. degrees in electrical engineering at Stanford University in 1948, 1949, and 1952, respectively.

From 1949 to 1951 Dr. Stewart was employed at the California Institute of Technology Jet Propulsion Laboratory and The Hughes Aircraft Co., Culver City, California. From 1952 to 1953, he was a research associate at the Stanford University Electronics Research Laboratory. Since 1953, Dr. Stewart has been assistant professor of electrical engineering at the University of Michigan and consultant to the Engineering Research Institute at the University.

Dr. Stewart is a member of Tau Beta Pi and Sigma Xi.

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R. P. Stone

R. P. Stone (S'40–A'42–M'46–SM'50) was born on April 10, 1918, in Columbus, Ohio. He received the B.E.E. degree in 1940 from the Ohio State University, the M.S. degree in 1941 from Purdue University, and the Ph.D. degree in electrical engineering from Princeton University in 1949.

Dr. Stone joined the Radio Corporation of America RCA Laboratories Division, in 1941, working on various problems in the field of electron tube research.

In addition to his I.R.E. activities he is a member of Sigma Xi.

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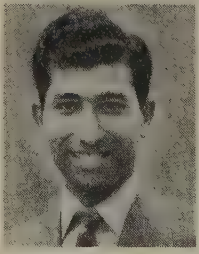

B. H. Wadia

B. H. Wadia (S'50–A'53) was born on October 1, 1927 in Surat, India, He received the degree of B.E. (Mechanical and Electrical) at the University of Bombay in 1948 and came to the United States for graduate work in Electrical Engineering at Stanford University, Stanford, California. After obtaining the M.S. degree in 1950, he worked as a Research Assistant at Stanford and obtained his Ph.D. degree in April, 1954.

Dr. Wadia now is a Research Associate at the Microwave Laboratory, Stanford University.

In addition to being a member of I.R.E., Dr. Wadia is a member of Tau Beta Pi and Sigma Xi.

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R. H. Wilcox

R. H. Wilcox (S'48–A'51) was born in Wooster, Ohio, on September 23, 1927. He received the B.S. degree in electrical engineering from Lafayette College in 1951.

Previous to this he was associated with the RCA Service Company and All American Airways as an electronic technician, and he served two years as an Aviation Electronic Technician in the U. S. Navy. Mr. Wilcox has been employed by the Naval Research Laboratory, Washington, D. C., since 1951, where he has engaged in research on noise generation and electronic analog devices for the computation of statistical signal parameters. At present he is associated with the Command Guidance Section of the Aviagation and Transistor Research Branch, Electronics Division.

Mr. Wilcox is an associate member of the A.I.E.E. and the U. S. Naval Institute, and a member of Tau Beta Pi.

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E. A. Wolff

E. A. Wolff (S'51) was born on October 31, 1929 in Chicago, Illinois. He received the B.S. degree in Electrical Engineering from the University of Illinois in 1951. In 1953 he received the M.S. degree in Electrical Engineering from the University of Maryland.

Since his graduation from Illinois, Mr. Wolff has been with the Wave Propagation Branch of the Naval Research Laboratory in Washington, D. C., where he has devoted his time to the development of experimental radar systems.

Beside his Senior Membership in I.R.E., Mr. Wolff is an Associate Member of the AIEE. He also holds membership in Eta Kappa Nu and Sigma Tau.

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IRE News and Radio Notes

FIRST U. S. AUTOMATION SHOW TO BE HELD

The world of the future, in which factories which presently operate with 2,000 employees will be manned by a few hundred, and in which the three-day work week will be standard, will be previewed at the nation's first automation show, November 21–22 at the Waldorf-Astoria Hotel, New York, N. Y.

Eighty companies in the field of control systems, components and services will be represented as exhibitors. The show will dramatize the enormous possibilities of computers and control "hardware" and also demonstrate precisely what is being accomplished today. Called the Automatic Control Equipment Show, the gathering will be held for executives and engineers on a top management level rather than for technical specialists or the general public.

Exhibitors will include instrument and computer manufacturers, firms which design and install complete automatic systems, and makers of the components comprising these systems. Additional information about the exhibit may be obtained from Strauss, Spigler and Kline Corporation, Finance Building, South Penn Square, Philadelphia, Pa.

CONFERENCE ON NUCLEAR ENGINEERING TO BE HELD IN 1955

An advance announcement of a 1955 Conference on Nuclear Engineering on April 27–29 at the University of California at Los Angeles, and an invitation to submit papers to be read at the gathering, have recently been issued by the Departments of Engineering on the Los Angeles and Berkeley campuses of the University.

The Conference, offered through University Extension and sponsored by the outstanding national engineering societies, will consist of three one-day sessions, each devoted to one of the following topics: Nuclear Reactor Power Extraction Systems and Auxiliaries (liquid metal systems, high pressure water systems, etc.); Nuclear System Dynamics (power system dynamic behavior, reactor, kinetics, instrumentation and control, etc.); and Radiation Sources for Industrial Applications.

Any person wishing to present a paper on one of these topics should contact Professor T. J. Connolly, Department of Engineering, University of California, Los Angeles 24, California. A description of the AEC-approved security-clearance procedure for papers will be mailed with notice of tentative acceptance of these proposed papers. Preprints of papers selected will be available at the Conference.

FIRST MEETING BOSTON CHAPTER MICROWAVE THEORY AND TECHNIQUES GROUP

The first official meeting of the Boston Chapter of the I.R.E. Professional Group on Microwave Theory and Techniques was held on June 3, 1954, at the Massachusetts Institute of Technology, Room 3–370. Fifty-three people were present.

The nominating committee, Merrill I. Skolnik, Sylvania, Boston; Leslie M. Vant, Diamond Microwave; and Daniel A. Lanciani, Microwave Associates, approved the following selected candidates for office: Chairman Theodore S. Saad, Sylvania, Woburn; Vice-Chairman Walter Rotman, Cambridge Research Center; Secretary-Treasurer Wilbur L. Pritchard, Raytheon, Newton.

The latter half of the meeting was taken up by a very interesting paper, "Ferrites in Waveguides," given by Dr. Robert H. Fox, of the M.I.T. Lincoln Laboratories.

Final Call For Papers

IRE NATIONAL CONVENTION MARCH 21–24, 1955

Prospective authors are requested to submit all of the following information:

(1) 100-word abstract in triplicate with title, name and address of author.
(2) 500-word summary in triplicate with title, name and address of author.
(3) Indicate the technical field in which your paper falls:

Aeronautical & Navigational Electronics
Antennas & Propagation
Audio
Broadcast & Television Receivers
Broadcast Transmission Systems
Circuit Theory
Communications Systems
Component Parts
Electron Devices
Electronic Computers
Engineering Management
Industrial Electronics
Information Theory
Instrumentation
Medical Electronics
Microwave Theory & Techniques
Nuclear Science
Production Techniques
Quality Control
Radio Telemetry & Remote Control
Ultrasonics Engineering
Vehicular Communications

Deadline for acceptance of papers: November 15, 1954

Address all material to:

Mr. J. Z. Millar, Chairman
1955 Technical Program Committee
Institute of Radio Engineers, Inc.
1 East 79 Street, New York 21, N. Y.

Calendar of COMING EVENTS

IRE Professional Group on Vehicular Communications Meeting, Rice Hotel, Houston, Texas, September 30–October 1

National Electronics Conference, Hotel Sherman, Chicago, Ill., October 4–6

IRE Professional Group on Nuclear Science Annual Conference, Sherman Hotel, Chicago, Ill., October 6–7

Symposium on Marine Communication and Navigation, Hotel Somerset, Boston, Mass., October 13–15

IRE-RETMA Radio Fall Meeting, Hotel Syracuse, Syracuse, N. Y., October 18–20

National Association of Educational Broadcasters 30th Annual Convention, Biltmore Hotel, New York, N. Y., October 27–30

IRE Baltimore Section-PGANE East Coast Conference on Airborne and Navigational Electronics, Sheraton-Belvedere Hotel, Baltimore, Md., November 4–5

IRE-PIB Microwave Symposium, Engineering Societies Auditorium, New York, N. Y., November 8–10

IRE-AIEE Conference on Electrical Techniques in Medicine and Biology, Morrison Hotel, Chicago, Ill., November 10–11

IRE Quality Control Symposium, Statler Hotel, New York, N. Y., November 12–13

Symposium on Fluctuation Phenomena in Microwave Sources, Western Union Auditorium, New York, N. Y., November 18–19

IRE Kansas City Section Annual Electronics Conference, Hotel President, Kansas City, Mo., November 18–19

IRE-AIEE-ACM Eastern Computer Conference, Bellevue-Stratford Hotel, Philadelphia, Pa., December 8–10

IRE-AIEE-NBS-URSI Conference on High Frequency Measurements, Hotel Statler, Washington, D. C., January 17–19

IRE National Convention, Waldorf-Astoria Hotel and Kingsbridge Armory, New York, N. Y., March 21–24

IRE Seventh Region Technical Conference, Hotel Westward Ho, Phoenix, Ariz., April 27–29

National Airborne Electronics Conference, Biltmore Hotel, Dayton, Ohio, May 9–11

AUTOMATION SYMPOSIUM IN PHILADELPHIA

Six weekly lectures on automation, sponsored by the Philadelphia Sections of the IRE and AIEE, October 10–November 20, will cover: Automation in the Automotive Industry; Electronic Auto-Assembly; Etched Circuits; Modular Approach to Mechanized Assembly; Process Control; and Automation in Chemical Plants.

For registration information, write the IRE Symposium on Automation, 1317 Spruce Street, Philadelphia 7, Pa.

COWAN AWARDED LAMME MEDAL

Frank A. Cowan (M'30–SM'43) assistant director of operations, Long Lines Department, American Telephone & Telegraph Company, and holder of 17 patents in the communications field, has recently been awarded the 1953 Lamme Gold Medal. Specifically, Mr. Cowan was cited "for his outstanding contributions to long distance communication and the development of modulating and transmission measuring apparatus of original design and application."

The Lamme Medal is awarded annually to a member of the American Institute of Electrical Engineers "who has shown meritorious achievement in the development of electrical apparatus or machinery." It has been awarded annually since 1928. In the interim some of the nation's top electrical engineers have received it; Mr. Cowan is the first in the field of telephone engineering to receive the medal.

Many of Mr. Cowan's inventions are widely used throughout the industry. Among them are bridge and ring type modulators and demodulators, used today in communications systems. Another of his inventions first built in his own workshop and now employed throughout the Bell System, is a direct reading telegraph transmission measuring set.

Mr. Cowan is the author of many technical papers on telephone subjects, and is a Fellow of the American Institute of Electrical Engineers.

WE STAND CORRECTED

John D. Ryder has brought the following correction to the attention of the editors. In his biography under IRE People, p. 1034, June, 1954 issue of the PROCEEDINGS OF THE I.R.E. it was stated that "On July 1, John D. Ryder will take over his new duties as Dean of the School of Electrical Engineering at Michigan State College, East Lansing, Mich."

Dr. Ryder's present position is Dean of the School of Engineering. The word "Electrical" should not have appeared in the above information.

SYMPOSIUM PROCEEDINGS ON AUTOMATIC PRODUCTION OF ELECTRONIC EQUIPMENT PUBLISHED

Proceedings of the Symposium on Automatic Production of Electronic Equipment held last April 19–20 in San Francisco, have been published by the Stanford Research Institute. The symposium was jointly sponsored by the Air Force and SRI.

The 119-page bound volume contains 17 papers and illustrations relating to the general aspects of automation, product design, construction techniques, materials, components and the design of automatic production lines; also a transcript of the symposium's concluding panel discussion.

For information write Public Relations Office, Stanford Research Institute, Stanford, Calif.

DR. TULLER AIR CRASH VICTIM

Dr. William G. Tuller of Falls Church, Virginia, died in the K.L.M. crash near Shannon Airport, Ireland, on Saturday, September 4, 1954.

PAPERS INVITED FOR SEMICONDUCTOR SYMPOSIUM

The third annual Semiconductor Symposium of the Electrochemical Society will be held this Spring, from May 2–5, at Cincinnati, Ohio. Papers are invited for presentation. Our early plans are to divide the sessions into two half-day sessions on semiconducting materials—elemental, alloys and compounds; a half-day session on surface controlled phenomena; and a half-day session on chemical process technology.

In each of the areas of interest we would like to present a balanced agenda, including one paper reviewing the state of each field, 30-minute presentations (including discussion time) of new information that can be scheduled by mid-winter 1954, and a number of shorter presentations of about 10 minutes' duration of the "late news" type.

Those wishing to present papers should notify the chairman, also providing the title of any proposed paper, as early as possible, and no later than November 15, 1954. Such information is needed for agenda planning purposes. An abstract of approximately 75 words should be submitted no later than January 15, 1955. This will be printed in the general program for the meeting. A second "extended abstract" of about 1,000 words should be submitted, along with brief pertinent data, illustrations, curves, etc., not later than February 1, 1955. This extended abstract is "printed, but not published" in a booklet entitled "Enlarged Abstracts of Papers Presented by the Electronics Division," and will be available at the meeting at cost.

The short "late news" presentations will necessarily be less publicized in advance of the meeting than are the longer papers. Nevertheless, they will be announced, to the division mailing list before the meeting.

All communications should be sent to Mr. F. J. Biondi, Chairman, at Bell Telephone Laboratories, Murray Hill, N. J.

Professional Group News

FIRST ANNIVERSARY OF THE PGUE

The PGUE Administrative Committee held its fourth meeting in New York City on June 25, 1954, on the occasion of the first anniversary of the PGUE. Three new members to the Administrative Committee were elected by the holdover members. The three were chosen from a slate of six, nominated by a Special Nominating Committee. The latter Committee consisted of David Arenberg (Arenberg Ultrasonic Laboratory), Donald Berlincourt (Brush Laboratories), and Walter G. Cady (California Institute of Technology).

The three new members elected to three year terms are as follows: Warren P. Mason, Bell Telephone Laboratories; Julia Herrick, Mayo Clinic; Karl Van Dyke, Wesleyan University.

The three retiring members of the Administrative Committee are: W. G. Cady, W. J. Mayo-Wells, and P. L. Smith.

Since the Group is just entering its second year, the Committee decided to re-elect the same officers for another year. They include Amor L. Lane, chairman; Morton D. Fagen, Vice-Chairman; Morris Kenny, Secretary. In addition, Julius Bernstein was selected to be Treasurer.

Likewise, the same Committee chairmen were retained: M. D. Fagen, Membership, and Oskar Mattiat, Papers Procurement. In addition, William J. Fry was appointed chairman of the Papers Study and Review Committee, Julius Bernstein taking over the Sectional-Chapters Activities Committee.

Dr. Oskar Mattiat (Brush Laboratories) is the Editor-in-Chief of the PGUE TRANSACTIONS (the first issue was published in July, 1954), with Donald Berlincourt nominated as Associate Editor.

The PGUE Administrative Committee formally approved the Group Constitution and By-Laws, to be circulated to the memship in the near future.

The meeting concluded by electing Francis X. Byrnes (USN Electronics Laboratories, San Diego) as chairman of the Ultrasonics panel at the August, 1954 WESCON.

DUNMORE RECEIVES AERONAUTICAL AWARD

In 1951 the Professional Group of Aeronautical and Navigational Electronics established an award for pioneers in the field of aeronautical and navigational electronics. The award this year was made to F. W. Dunmore. Mr. Dunmore was born January 24, 1891, at Haverhill, Mass. He became a commercial radio operator in 1911–1913; received the B.S. degree from Pennsylvania State College in 1915; was a student of engineering at General Electric Company from 1915–1917; joined the Research Department, American Radio and Research Corporation in 1918 for one year; and has since been a staff member of the National Bureau of Standards, Washington, D. C.

Many new aids, such as the omnidirectional radio range, have been installed in the United States through the NBS since the war. This facility is now beginning to come into routine use, and by far the greatest number of miles are flown in the United States by means of the four-course radio range system. The first unit of this system was installed in 1929. Work on it was started in the NBS, during 1922 and 1923, by F. H. Engelman and F. W. Dunmore. Thus, Mr. Dunmore played a very important part in the development of the standard United States air navigational aid.

Mr. Dunmore's work, however, does not cease with the development of the four-course radio range. In 1930 he began work on the fixed-beam low-approach system, commonly referred to as ILS. At the outbreak of the war, the Civil Aeronautics Administration had half a dozen ILS systems in operation, and were preparing to install this equipment throughout the country. During the war, the CAA system was modified by the Air Force and some 500 equipments were constructed for use of Air Corps aircraft. Subsequent to the war, the ILS system was standardized by the International Civil Aviation Organization for use throughout the world. Thus Dunmore's work is guiding aircraft to safe landing in low visibility throughout the world.

The Professional Group on Aeronautical and Navigational Electronics is happy indeed to make the Pioneer Award to Mr. Dunmore. Without a doubt he is one of the most outstanding radio engineers working today who have contributed very substantially to aeronautical electronic science.

CUMULATIVE INDEX AVAILABLE

The Cumulative Index for the PROCEEDINGS OF THE I.R.E., TRANSACTIONS of the I.R.E. Professional Groups, and the CONVENTION RECORD OF THE I.R.E., for the period 1948–1953, is now available. The Index will be sold to members at a price of $1.00, and to nonmembers at a price of $3.00. To receive a copy of the Index, send orders with payments to The Institute of Radio Engineers, 1 East 79 Street, New York 21, N. Y.

Obituaries

Paul M. Gunzbourg (M'43–SM'43) consulting engineer, died recently.

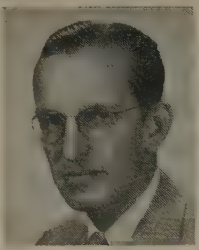
A graduate electrical engineer of the Polytechnical Institute of St. Petersburg, Russia, Mr. Gunzbourg served in World War I as a member of the Supreme Council and Defense of the Imperial Territory, and until the Russian revolution, was chief engineer and manager of the Russian Siemens Company under goverment sequester. From 1926 to 1940, he was president and general manager of the Belgian and French Siemens Companies. At the beginning of World War II he represented the British government, His Majesty's Office of Works, in France.

P. M. GUNZBOURG

He arrived in the United States in 1940 and in his capacity as consultant to the director of intelligence of the U. S. Army Service Forces, at the end of the war was awarded a Certificate of Appreciation for his "patriotic services in a position of trust and responsibility, whose advice on matters pertaining to French electrical, chemical, and mineral assets, as well as on problems encountered in connection with occupation, relief, and rehabilitation, was of invaluable assistance to the department."

Mr. Gunzbourg was a licensed professional engineer in New York State, and a member of the AIEE, The French Society of Electrical Engineers, the Belgian Society of Electrical Engineers, the International Conference of Power Grids in Paris, France, and the National Society of Professional Engineers, New York Chapter.

Lawrence V. Wells (A'29–SM'47) died recently at his home in Kalamazoo, Mich.


L. V. WELLS

Mr. Wells was born in Clyde, Ohio, in 1906. He studied electrical engineering at Michigan State College, Lansing, and later joined the staff of radio station WREO, Lansing, during construction. He continued with WREO until 1927, when the license was surrendered and broadcasting abandoned. With a partner he operated a radio business until joining the engineering staff at Wilcox-Gay at Charlotte, Mich. He became chief engineer and served Wilcox-Gay in that capacity until 1947, when he resigned to take charge of engineering at Allen Electric Equipment Company, Kalamazoo. He entered selling in 1953, joining Haggerty Sales Company, Ferndale, Mich. After a brief period, a partnership, Haggerty-Wells Company was formed, and a branch at Kalamazoo set up. He was on the consulting staff of several prominent electrical and electronic concerns, and was widely known for his work in mass-produced recorders, FM, AM, TV, electronic musical instruments, telemetering audio, and related equipment.

Jerome A. Beranek (S'41–A'42–M'46–SM'48) Manager of CBS Radio Technical Operations in Hollywood, died recently of a heart attack at the age of 38.

Mr. Beranek was born in Saskatoon, Saskatchewan, Can. He had been with CBS Radio in Hollywood since his graduation from the University of Southern California in 1940, with the exception of three war years spent in the Bureau of War Research of Columbia University, New York.

Cornelius G. Brennecke (A'36–M'46–F'51), head of the North Carolina State College Department of Electrical Engineering, died recently. He was 47 years old.

C. G. BRENNECKE

Dr. Brennecke, a native of New York, received the degrees of A.B., B.S., and E.E. from Columbia University. His graduate work was done at New York University, where he received the degree of Ph.D. in 1936.

After several years with the engineering department of RCA, he accepted a fellowship in physics at New York University, leaving there in 1936 to join the engineering faculty of the University of Toledo. In 1942 he became an associate professor of electrical engineering at Lehigh University. In 1949 he was appointed North Carolina State College representative on the Oak Ridge Institute of Nuclear Studies. His investigations in atomic physics and in the properties of dielectrics have been published.

Dr. Brennecke helped to organize the North Carolina-Virginia Section of the IRE, and served as its chairman in 1948, and as vice-chairman in 1949–50.

SEVENTH ANNUAL CONFERENCE ON ELECTRICAL TECHNIQUES IN MEDICINE AND BIOLOGY

SPONSORED BY THE AMERICAN INSTITUTE OF ELECTRICAL ENGINEERS, THE INSTITUTE OF RADIO ENGINEERS, AND THE INSTRUMENT SOCIETY OF AMERICA

Morrison Hotel, Chicago, Ill., November 10–12

Wednesday morning November 10

CIRCULATION AND CARDIOLOGY

Chairman, Dr. Samuel A. Talbot

1. "Normalization of Vector EKG Axes," Dr. Otto Schmitt, the University of Minnesota.
2. "Computer Technics in Rapid Determination of Certain EKG Functions," Dr. Stanley Briller, Bellevue Hospital.
3. "Synthesis of Electrocardiograph Leads," Dr. Richard McFee, Bell Telephone Laboratories, Whippany, N. J., and Dr. Franklin Johnston, University of Michigan Medical School.
4. "Determination of the Electrical Center of Ventricular Depolarization in the Human Heart," Dr. Ernest Frank, Moore School of Electrical Engineering, University of Pennsylvania.
5. "Heart Cell Potential," Dr. E. E. Suckling, New York State University.

Wednesday afternoon November 10

ELECTRICAL PROPERTIES OF BIOLOGICAL MATERIALS

Chairman, Dr. Otto Schmitt

1. The Electrical Impedance of a Human Body Segment as Related to Impedance Plethysmography," Dr. Paul Albro, Hastings Instrument Co.
2. "Electro-ionics of Nerve Action," Dr. Kenneth S. Cole, Naval Medical Research Institute, Bethesda, Md.
3. "The Application of the Electric Impedance Method in the Study of the Normal and Hemolyzed Red Blood Cell," Dr. Hugo Fricke, Walter B. James Laboratory for Biophysics, Cold Spring Harbor, N. Y.
4. "Capacity of Erythrocyte Ghose Membrane Measured over a Wide Frequency Range," Drs. Herman P. Schwan, E. L. Carstensen, and K. Li, Moore School of Electrical Engineering, University of Pennsylvania.

Thursday afternoon November 11

X-RAYS AND INSTRUMENTATION

1. "Spectral Phonocardiography," Dr. G. N. Webb, Johns Hopkins Hospital.
2. "New Developments in the Electronics Fluoroscope of the Television Type," Mr. Ralph Sturm and Dr. Russell Morgan, Johns Hopkins Hospital.
3. "Effects of X-Rays on Erythrocytes," Dr. Otto Blüh, University of British Columbia, Vancouver, Canada.
4. "Xerography," Dr. John Roach, Albany Medical School.
5. Progress in X-Ray Movies," E. L. Webb, Westinghouse Electric Corp., Baltimore, Md.

A dinner and evening meeting are scheduled for November 10th. The guest speaker will be Dr. T. E. Allibone, Director of Research, of the Associated Electrical Industries of Great Britain. Dr. Allibone is a Fellow of the Royal Society and known internationally for his work in fields pertinent to the Conference.

Two laboratory field trips have been arranged. Registrants for the Conference will be given an opportunity to visit the Argonne Cancer Research Hospital on the morning of November 11th. On the morning of November 12th, they will have an opportunity to visit appropriate portions of the Argonne National Laboratory. Both these installations, operated by the University of Chicago for the United States Atomic Energy Commission, are engaged in important studies of the effects of radiation on living systems.

The visit to the Argonne Cancer Hospital on the morning of the 11th will include inspection of the 50 mev microwave linear accelerator, the colbalt-60 rotational therapy unit, the thulium diagnostic unit, etc.

The visit to the Argonne National Laboratory will take place on the morning of November 12th. Transportation to and from this laboratory will be available and can be arranged for at the Conference registration desk. Maps and directions will be available for those having their own transportation. The visit will be planned to permit departure from Chicago by later afternoon train or plane.

Visitors to the Argonne National Laboratory will be asked to register for the tour before 3:00 P.M. of November 11th. Citizens of countries other than the United States must give three weeks notice through the Conference Chairman so that admission to the Argonne National Laboratory may be arranged.

Conference Chairman, E. D. Trout, 4855 Electric Ave., Milwaukee 1, Wis.

RADIO FALL MEETING PROGRAM

SPONSORED BY IRE, RETMA, AND RTMA OF CANADA
HOTEL SYRACUSE, SYRACUSE, N. Y. OCTOBER 18–20

Monday, 9:30 A.M., October 18

General Meeting

(Sponsored by the RETMA Engineering Department)

W. R. G. Baker, *presiding*

The program of this session will be comprised of reports by the Chairmen of the Sections and Staff Committees of the RETMA Engineering Department of activities and plans of their respective groups.

2:00 P.M.

Reliability Control Session

(Arranged by IRE Professional Group on Quality Control)

J. R. Steen, *presiding*

"Contributing Factors to Component Parts Reliability and Extended Service," J. A. Goetz, International Business Machines Corp.

"More Protection From Small Samples Through Better Planning," B. P. Goldsmith, Raytheon Manufacturing Co.

"Cathode Interface Impedance Desimplified," H. B. Frost, Bell Telphone Labs.

"Statistical Design—A Means to Better Products of Lower Cost," R. C. Miles, Airborne Instruments Lab.

Tuesday, 9:00 A.M., October 19

Television Receiver Session

(Arranged by the IRE Professional Group on Broadcast and Television Receivers)

E. I. Anderson, *presiding*

"Optimum Crystal Mixer Operation—the 1N82 Crystal," S. Deutsch, Polytechnic Res. & Dev. Co., Inc.

"A UHF-VHF Television Tuner Using Pencil Tubes," W. A. Harris and J. J. Thompson, R.C.A.

"Color Purity in Ungated Sequential Displays," G. S. Ley, Westhinghouse Elec. Corp.

"The Practical Aspects of Color Subcarrier Synchronization Problem," W. J. Gruen, General Electric Co.

"A Simple Method of Phase Compensation of Video Delay Lines," D. A. Gillen, Admiral Corp.

2:00 P.M.

General Receiver Session

(Arranged by the IRE Professional Group on Broadcast and Television Receivers)

D. W. Pugsley, *presiding*

"New Converter for Broadcast Receivers," D. E. Sunstein, Philco Corp.

"The Second Detector—A Determinant of Fringe-Area Performance," J. E. Bridges, Zenith Radio Corp.

"Dynamic Diode Limiters in FM Demodulator Circuits," F. Mural, Philco Corp.

"Automatic Gain Control of Transistor Amplifiers," W. F. Chow and A. P. Stern, General Electric Co.

Wednesday, 9:20 A.M., October 20

Electric Devices Session I

(Arranged by the IRE Professional Group on Electron Devices)

G. A. Espersen, *presiding*

"The Application of Frame Grids to Receiving Tubes," S. L. Pawlikowski, Sylvania Elec. Prods., Inc.

"A High-Voltage Regulator Tube for Color-Television Receivers," R. E. Byram, RCA.

"The Noise Figure of Grid Controlled Electron Tubes," H. Rothe, Telefunken, Ulm/Donau, Germany.

"Effects of Lead Inductance on the Performance of Miniature Tubes in the UHF Television Band," R. N. Peterson and W. A. Harris, RCA.

"A New Approach to Converter Tube Design," R. J. Bisso, Sylvania Elec. Prods., Inc.

2:00 P.M.

Electron Devices Session II

(Arranged by the IRE Professional Group on Electron Devices)

J. S. Saby, *presiding*

"Investigation of Noise in Audio-Frequency Amplifiers Using Junction Transistors," P. L. Bargellini and M. B. Herscher, RCA.

"Class B Operation of Audio-Frequency Junction Transistors," K. E. Loofbourrow, RCA.

"Raytheon Ck 721 Transistor. A Service Reliability Report," F. M. Dukat, Raytheon Manufacturing Co.

"Transistor Reliability," C. L. Rouault, General Electric Co.

NATIONAL SYMPOSIUM OF QUALITY CONTROL AND RELIABILITY IN ELECTRONICS

SPONSORED BY PROFESSIONAL GROUP ON QUALITY CONTROL OF THE IRE AND ELECTRONIC TECHNICAL COMMITTEE OF THE ASQC

Statler Hotel, New York, N. Y.—November 12 and 13, 1954

SESSION I

Friday, November 12, 1954

QUALITY CONTROL IN ENGINEERING AND MANUFACTURE

Moderator: To Be Announced

9:00–9:45 A.M.

1. "Statistically Designed Experiment of the Factorial Type Applied to Point-Contact Transistors," M. Eder, R. Warner, and F. Keene—Bell Telephone Labs.

9:45–10:30 A.M.

2. "The Bendix Radio Vender Rating System," R. Fitzgibbons—Bendix.

10:30–10:45 A.M.

Intermission

10:45–11:30 A.M.

3. "Sampling Procedures on Finished Chassis and Equipment," H. Knapp—Federal Telephone & Radio.

11:30–12:15 A.M.

4. "Some Aspects of Quality Control in Computer Tube Applications," W. Hall—IBM.

SESSION II

QUALITY CONTROL AND ITS RELATION TO RELIABILITY

1:30–2:15 P.M.

1. "Technical Publications' Contribution to Reliability," J. Wight—McGraw-Hill Book Co., Inc.

2:15–3:00 P.M.

2. "Managing a Quality Program," H. Newton—Bendix Aviation, Towson, Md.

3:00–3:45 P.M.

3. "An Integrated Program for Reliability Improvement," G. M. Armour—General Electric Co., Syracuse, N. Y.

3:45–4:00 P.M.

Intermission

4:00–4:45 P.M.

4. "Electronic Product Reliability Through Quality Control," C. Ryerson—RCA.

4:45–5:30 P.M.

5. "Feedback Approach to Reliability," D. Hill and H. D. Voegtlen—Hughes Aircraft Co.

SESSION III

Saturday, November 13, 1954

MILITARY ASPECTS OF RELIABILITY

9:00–9:45 A.M.

1. "The Navy Reliability Program and the Designer," E. J. Nucci—Navy Dept., Bureau of Ships, Washington, D. C.

9:45–10:30 A.M.

2. "Development of Techniques for Reliability Measurement and Prediction," V. Harris and M. Tall—Vitro Corp.

10:30–11:15 A.M.

3. "Concepts of Failure Reporting and Its Relation to Reliability," C. E. McLaughlin—Wright Patterson AF Dev. Center.

11:15 A.M.–12:00 N.

4. "Techniques in Putting Failure Data to Work for Management," F. A. Hadden and L. W. Sepmeyer—Rand Corp.

12:15–2:30 P.M.

Luncheon Session

Speaker: To Be Announced

SESSION IV

AIRBORNE RELIABILITY

2:45–3:30 P.M.

1. "Designing Airborne Electronic Equipment for Improved Reliability," R. N. White, Trans-World Airlines, Inc., Kansas City, Mo.

3:30–4:15 P.M.

2. "The ARINC Military Electron Tube Project—Its History, Philosophy and Major Findings," E. Morse—Aeronautical Radio, Inc.

4:15–5:00 P.M.

3. "Development, Implementation, and Utilization of the '*K*' System Consolidated Malfunction Report," F. Tonnies—Sperry Gyroscope Co.

SYMPOSIUM ON FLUCTUATION PHENOMENA IN MICROWAVE SOURCES

November 18–19, New York, N. Y.

A Symposium on Fluctuation Phenomena in Microwave Sources will be held at the Western Union Auditorium, New York City, on November 18–19, 1954, under the joint sponsorship of the IRE Professional Group on Electron Devices and the Advisory Group on Electron Tubes.

The symposium has been organized with the following objectives in mind: (a) To acquaint conferees with the noise requirements of various types of microwave equipment; (b) to formulate plans for standardization of definitions of noise and noise measurements; and (c) to acquaint microwave engineers with the types and relative amounts of noise to be expected from various microwave sources.

It is planned that all papers will be published after the symposium in the *Transactions* of the Professional Group on Electron Devices.

No registration fee will be charged for the symposium. Those who wish to attend should write to Col. H. W. Serig, Chairman, Invitations Committee, Advisory Group on Electron Tubes, 346 Broadway (8th floor), New York, N. Y.

The program and abstracts of papers are as follows:

SESSION I

NOISE REQUIREMENTS OF MICROWAVE SYSTEMS

Session Organizer: W. M. Gottschalk, Raytheon Mfg. Co., Waltham, Mass.

Session Chairman: W. C. Brown, Raytheon Mfg. Co., Waltham, Mass.

Session Periods: 10:00 A.M.–12:15 P.M. Thursday, November 18, 1954.

PAPERS

Relation of Noise in Microwave Sources to System Requirements—J. B. Wiesner, Mass. Inst. of Tech., Cambridge, Mass.

A brief survey is given of the noise characteristics of the common microwave sources, including triode oscillators, magnetrons, and klystrons; followed by a discussion of the limitations which such electrical noises impose on communications and radar systems. The paper will include a discussion of the effects of additive noise, amplitude-modulation noise, frequency-modulation noise and flicker noise.

The Influence of Noisy Components on the Sensitivity of Microwave Receivers—W. L. Pritchard and K. T. Larkin, Raytheon Mfg. Co., Waltham, Mass.

The ultimate limitation on receiver sensitivity is random noise. After establishing basic quantities such as bandwidth, type of modulation, etc., the remaining problem is to come as close as possible to the minimum theoretical noise level.

We define noise figure and apply it first to cascaded networks and then to a superheterodyne receiver using a crystal converter. The following basic equation is derived:

$$F = L_R L_m [(F_{\text{if}} - 1) + (t_m - 1) + 1].$$

We discuss the parameters in the above equation and show that, while these are often considered to be the published values for the crystal and if amplifier tube, they are actually significant functions of the associated rf and if circuitry.

We discuss the effects of extraneous frequency generation and rf source impedance on conversion loss and noise temperature.

We next consider the sources of noise in an if amplifier and show that there is an optimum if source impedance. The existence of an optimum if frequency is established and its dependence on the crystal and tube parameters is shown. We emphasize the criteria for correctly connecting the if amplifier to the mixer.

We show how local oscillator noise contributed to noise figure and how it varies with if and carrier frequencies. Experimental data are presented showing the considerable magnitude of the effect.

Stabilization of Microwave Oscillators—E. J. Shelton, Raytheon Mfg. Co., Waltham, Mass.

Frequency stabilization of microwave oscillators by the use of resonant cavities will be reviewed. The benefits of using a damping resistor to suppress unstabilized and unwanted modes of operation are included. During the course of studying system performance of a particular microwave system, it was determined that unintentional stabilization was seriously effecting operation. This unintentional stabilization can be either beneficial or detrimental in a particular instance, but if not recognized and controlled the general result is adverse. Theory and experimental results are to be presented along with possible means of making use of the advantages of this effect to decrease frequency variations.

SESSION II

REPRESENTATION AND MEASUREMENT OF FLUCTUATION SIGNALS

Session Organizers: H. D. Arnett, Naval Research Lab., Washington, D. C.; R. T. Young, Diamond Ordnance Lab., Washington, D. C.

Session Chairman: A. J. Ruhlig, Electron Tube Branch, Naval Research Lab., Washington, D. C.

Session Period: 1:30 P.M.–5:00 P.M., Thursday, November 18, 1954.

PAPERS

Noise Measurements of Microwave Local Oscillators—By Dr. R. Mueller, presented by H. J. Hersh, Evans Signal Lab., Belmar, N. J.

A method for measuring spectral distribution of amplitude fluctuations and frequency fluctuations in microwave oscillators is described. The method provides for evaluation of amplitude fluctuations in terms of signal-to-noise ratio; frequency fluctuations in terms of two parameters: (a) rms deviation, and (b) rms rate of change of deviation. In addition, the method permits estimation of the correlation between these two noise components.

Some typical results of such measurements are described, and certain implications of the effects of system parameters on interpretation of these results are considered.

For a specific system, two measurements are shown to be sufficient to determine the effect of frequency noise on system performance.

Microwave Oscillator Noise Spectrum Measurements—G. C. Dalman and A. S. Rhoads, Jr., Sperry Gyroscope Co., Great Neck, L. I.

Measurements, equipment and techniques have been developed for determining the noise spectrum of a microwave oscillator. In this method the oscillator output is passed through a microwave filter which sharply attenuates the power at the oscillator frequency but passes the noise frequencies. A noise-free local oscillator signal is added to the filter output and the oscillator noise spectrum is measured using a calibrated microwave receiver. A second microwave rejection filter is used in the measurements to eliminate one of the receiver responses.

A detailed description of the equipment and past techniques will be presented. Experimental results obtained on two typical 9 kmc. reflex klystrons operating at the center of the mode and electronically tuned to the half-power points, will also be described.

Theory of Phenomenological Models and Direct Measurements of the Fluctuating Output of CW Magnetrons—D. Middleton, Consultant, Research Div., Raytheon Manufacturing Co., Waltham, Mass.

A complete macroscopic theory yielding the significant parameters which describe the coherent and incoherent output of a cw magnetron (e.g. modulation and "pushing" factors, bandwidths, carrier and noise levels, modulation indexes, etc.) is outlined. The theory on which measurement of these descriptive parameters depends, is also described, as it is intimately related to the interpretation of the data from which the defining parameters of the phenomenological model are obtained. It is shown that the most satisfactory general model of cw operation is that of a carrier or primary oscillation simultaneously, amplitude and angle-modulated by a band of shot or "primary" noise, and accompanied by a background noise essentially incoherent with the modulated carrier. The chief effect is produced by the effects of angle-modulation, which even for relatively noisy tubes, occurs with a very low-modulation index; the amplitude-modulation, while relatively weak, cannot be ignored, and must be included in the analysis. A technique of direct measurement, using a waveguide filter and crystal detector for input waves of the above type, is briefly described, and it is shown how the descriptive parameters of the magnetron model may be observed experimentally. Particular attention is paid to the approximations and assumptions involved, and the theory is illustrated with a number of representative data from a standard magnetron; applications of this general approach to other microwave sources is briefly considered.

Direct Detection Measurements of the Output of CW Magnetrons—W. M. Gottschalk, Research Div., Raytheon Manufacturing Co., Waltham, Mass.

Apparatus will be described by means of which measurements have been made on the noise characteristics of cw magnetrons (not modulated). The resulting measurements show that the magnetron can be represented as an oscillator with the simultaneous amplitude and angle modulation, with a predominance of the latter. Variations of characteristics of the measuring equipment have disclosed certain properties. Numerical results on several tube types will be given and correlation with operating parameters will be shown.

Development of a Low-Noise *X*-Band CW Klystron Power Oscillator—R. A. La Plante, Philips Laboratories, Inc., Irvington-on-Hudson, N. Y.

A cw klystron oscillator for *X*-band made by Philips Laboratories delivers more than 100 watts at 9,600 mc/seconds. Using this tube as a prototype, alterations have been made in the tube design to reduce the frequency fluctuation of the output due to microphonism. The frequency fluctuations of the output of these tubes are expressed by the rms frequency deviation of the output, and a system used for the measurement of this quantity is described. Comparisons of the rms frequency deviations of conventional and low-noise designs of the Philips tubes are given.

The Measurement and Sources of Short Time Angular Instabilities—B. R. Mayo, H. H. Grimm, and J. K. Records, General Electric Co., Syracuse, N. Y.

The residual phase and frequency variations associated with highly stable microwave sources is discussed in this paper. The type of frequency and phase variations to be expected are listed and classified.

Measurement methods capable of detecting very small phase and frequency variations are described. The sensitivities obtainable are given. An objective of this work has been to find methods which are not too complex, but have adequate sensitivity for facilitating the development of sources with unusually good short time stability. Short time frequency deviations of one part in 10^{10} are easily observed. However, microwave sources tested so far never show less than 1 part in 10 degrees frequency deviation in periods of the order of 10^{-2} seconds.

Phase deviations of 0.1 degree can be observed. Equipment under investigation ordinarily shows 0.5 degree to 5 degrees phase shift.

Analysis is given relating the observed instabilities to some MTI problems.

A Technique for Measuring FM Noise in Microwave Oscillators—E. F. McClain and W. R. Ferris, Naval Research Lab., Washington 25, D. C.

Previous methods of analyzing the low-frequency noise spectrum of microwave oscillators have employed the usual scanning-type spectrum analyzer or an arrangement of fixed-tuned filters. The visual presentation of the spectrum analyzer makes high accuracy difficult, while a series of filters does not provide the necessary frequency discrimination. The present system makes use of a chart recorder to present noise level versus frequency as a continuous plot from 0 to 16,000 cycles.

National Bureau of Standards Noise Comparator—Dr. Charles Greenhow, Nat. Bur. Stand., Washington, D. C.

This paper is concerned with the measurements of the power spectrum of fluorescent tubes in the frequency range between 6–10 kmc. It has two main purposes: to discuss the experimental set-up used for calibrating fluorescent tubes to be used as secondary noise standards; and to study experimentally the law of available noise power of a discharge tube.

SESSION III-A

NOISE GENERATION IN MICROWAVE SOURCES

Session Organizers: C. F. Quate, Bell Telephone Labs., Murray Hill, N. J.; J. Babakian, Central Sales & Mfg. Corp., Denville, N. J.; A. G. Peifer, Federal Telecommun. Lab., Nutley, N. J.

Session Chairman: A. K. Wing, Federal Telecommun. Lab., Nutley, N. J.

Session Period: 9:15 A.M.–12:15 P.M., Friday, November 19, 1954.

PAPERS

General Sources of Noise in Microwave Tubes—J. R. Pierce, Bell Telephone Labs., Murray Hill, N. J.

This paper will discuss various sources of noise which are found in microwave amplifiers. The noise characteristics of the beams used in such amplifiers will be treated and used as a basis to predict the amount of noise output which can be expected from a given tube. Also, the various methods of reducing this noise output and a summary of the present state of low noise amplifiers will be presented.

Noise Measurements on Long Electron Beams—L. D. Smullin, Mass. Inst. Tech., Res. Lab. of Electronics, Cambridge, Mass.

Measurements of the microwave noise have been made on electron beams formed by several different guns. These have included parallel beam guns with cathode emersed in the magnetic field; and converging flow guns with cathode outside the magnetic field.

In general, two regions of behavior have been noted: one in which the noise current density varies sinusoidally with distance, and the other in which the noise current increases rapidly with distance. The latter behavior has been called the growing noise wave. The way in which these two different noise regimes respond to various parameters will be described.

Noise Characteristics of Carcinotrons—R. Warnecke, Compagnie Generale de la Telegraphie Sans Fils, Paris, France.

(No abstract available.)

The Small Signal Performance of the 416B Planar Triode between 60 and 4,000 mc—L. H. Von Ohlson, Bell Telephone Labs., c/o Western Electric Co., Allentown, Pa.

In view of the interest in the application of planar triodes as broadband, small signal amplifiers, a study of the performance of the 416B triode has been made between 60 and 4,000 mc. It has been found that this tube, having a gain-band product of 1,200 and an easily obtainable gain of 10 db at the higher frequencies, is capable of operation with a noise figure making it attractive for low level operation. Typical high frequency operation of the tube and a summary of the measurements made over a wide band of frequency and their agreement with theory is discussed.

Measurements of Traveling-Wave Tube Noise Figure—G. E. St. John, Bell Telephone Labs., Murray Hill, N. J.

The general structure of the tube is shown and a brief description of the electrode disposition given.

Variation of noise figure is described as a function of the first anode voltage with beam current as a parameter. The independence of current is pointed out.

A plot is shown of noise figure as a function of the perveance of the cathode to first-anode region and the possible significance of the beam shape in this region is described.

The variation of noise figure with frequency is described and some results are given on the effect of helix size.

Noise Characteristics of a Voltage Tunable Magnetron—J. A. Boyd, Univ. of Michigan, Ann Arbor, Mich.

The Michigan voltage-tunable magnetron is being studied to determine the noise characteristics of this tube. Amplitude of the noise as a function of frequency from the carrier is being investigated. Preliminary results indicate that the noise power is of the order of 100 db below the carrier at 30 mc from the carrier frequency. The tube is being used as the local oscillator in a microwave receiver and the noise contribution of this tube to the receiver output is being evaluated. An attempt will be made to correlate the experimental results with theoretical predictions.

Measurement and Analysis of Triode Noise—W. A. Harris, Tube Div., RCA, Harrison, N. J.

Measurements of noise from triode amplifiers are reported, under conditions which permit description of the noise as if derived from two generators. The mean-square amplitudes and the correlation coefficient for these generators are given for several tube types, over a range of operating conditions, for frequencies between 500 and 900 mc. The results are discussed in terms of their relation to triode theory.

SESSION III-B

THEORY OF NOISE GENERATION IN MICROWAVE SOURCES

Session Organizers: C. F. Quate, Bell Telephone Labs., Murray Hill, N. J.; J. Babakian, Central Sales & Mfg. Corp., Denville, N. J., A. G. Peifer, Federal Telecommun. Lab., Nutley, N. J.

Session Chairman: A. K. Wing, Federal Telecommun. Lab., Nutley, N. J.

Session Period: 1:30 P.M.–2:45 P.M., Friday, November 19, 1954.

PAPERS

Space Charge Waves on an Accelerating Stream of Uniformly Charged Square Laminae—W. E. Danielson, Bell Telephone Labs., Murray Hill, N. J.

Using an electronic model which consists of a stream of uniformly charged laminae accelerated in an electron gun, ac velocity and current at the anode will be given in terms of the corresponding quantities at the potential minimum. The physical principles on which the calculations are based will be summarized, and possible application of the results to traveling-wave tube noise theory will be outlined.

Observations on Ion Oscillations in a Cylindrical Beam Tetrode Under Hard Vacuum Conditions—W. E. Waters, Jr., Diamond Ordnance Fuze Labs., Connecticut Ave. at Van Ness St., N.W., Washington 25, D. C.

Observations have been made on ion oscillations occurring in ion traps in a number of tubes, operating over a wide range of voltage and current. Several geometries and electrode surfaces have been used; pressures have ranged from 5×10^{-8} mm Hg up to about 10^{-6} mm. Three distinct types of disturbances have been observed, all having fundamental frequencies in the range of 0.5 to 3 mc/seconds. Secondary electrons have been found to have a large influence on the strength of these oscillations, but only a small influence on the frequencies. The relation between the phenomena in the cylindrical-beam tetrode and sidebands occurring in the spectrum of a dc-operated reflex klystron has been examined; it seems clear that the ion oscillation produces simultaneous AM and FM of the klystron. A number of theoretical models have been examined, but none of them account for all the observations.

Noise Phenomena in the Region of the Potential Minimum—J. R. Whinnery, Electronics Res. Lab., Univ. of Calif., Berkeley, Calif.

Discussions are given on the importance of knowing noise behavior in the region of the potential minimum. Three attacks on this problem are described, a largely physical analysis, a series solution of the Liouville equation and other analytical attacks on this equation. Some of the phenomena yielded by these attacks are described with the serious problems which now prevent completing the solution.

Noise in Electron Beams—H. A. Haus, Mass. Inst. of Tech., Res. Lab. of Electronics, Cambridge, Mass.

An analysis based on the one-dimensional, single velocity assumptions leads to the result that all microwave amplifiers of the beam type (such as traveling-wave tubes, velocity jump amplifiers, klystrons, etc.) have a minimum obtainable noise figure of the form

$$F_{\min} = 1 + \frac{2\pi}{kT} K(S_0 - \pi_0) \qquad (1)$$

where S_0 and π_0 are parameters of the beam noise expressed in terms of the noise fluctuations at a cross section slightly beyond the potential minimum in the electron gun, T is the temperature of the circuit, and K is a constant of the particular structure employed. Expression (1) holds for any optimized noise reducing scheme. It includes, for instance, a prehelix which is excited by the noise in the beam and feeds into the main helix of a traveling-wave tube in an attempt to cancel part of the noise in the beam. The dependence of the constant K upon the particular structures used will be discussed.

The Signal to Noise Ratio in the M Carcinotron—O. Doehler and G. Convert

In this paper measurements are represented concerning the signal to noise ratio in the M type Carcinotron. The method of measurements is described and the experimental results are given. The signal to noise ratio is of the order of 3.10^{-13}/cc/sec for a few kc/su from the oscillating frequency, and 3.10^{-14}/cc/sec for 10 Mc/s from the oscillating frequency.

The signal to noise ratio increases approximately linearly with anode current.

The analysis of the sole current indicates, that this current is detected noise and it is assumed that growing space charge waves (diocotron effect) are the origin of the noise in the M type Carcinotron.

Physical Mechanism of Noise Generation in Magnetrons—C. G. Lehr and A. L. Collins, Res. Div., Raytheon Mfg. Co., Waltham, Mass.

Measured space-charge admittance values are used to calculate the following quantities associated with a traveling-wave cw magnetron: AM and FM noise; fluctuating anode current and voltage; and fluctuating rf voltage. All of these calculated quantities, except the rf voltage, are compared with measured values.

An insight into the physical mechanism of magnetron noise generation is obtained from relations between the space-charge admittance values and the electron density on the one hand, and the rf voltage fluctuation and the electron temperature on the other.

SESSION IV

SUMMARY AND DISCUSSION

Organizer and Session Chairman: W. H. Huggins
Session Period: 3:00 P.M.–5 P.M., Friday, November 19, 1954.
Summary of Important Points of Papers—W. H. Huggins, 15–30 minutes
Panel Discussion—30–60 minutes
Open Discussion—Remaining time.

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20.5 Definitions Coordinating

M. W. Baldwin, Jr., *Chairman*

P. S. Carter
J. G. Kreer, Jr.
E. A. Laport

20.8 Basic Terms

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W. Sichak

2.3 Antennas and Waveguides West Coast Subcommittee

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R. S. Elliott
S. Kerber
T. Moreno
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V. Twersky
R. S. Wehner

2.4 Waveguide and Waveguide Component Measurements

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P. A. Loth
A. A. Oliner
W. E. Waller

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4.2 Linear Lumped-Constant Passive Circuits

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4.3 Circuit Topology

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4.4 Linear Varying-Parameter and Non-Linear Circuits

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4.5 Time-Domain Network Analysis and Synthesis

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J. G. Linvill
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4.7 Linear Active Circuits Including Networks with Feedback and Servomechanism Circuits

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J. G. Linvill
J. M. Manley
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4.8 Circuit Components

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4.9 Fundamental Quantities

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J. D. Ryder

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7.2 Cathode-Ray and Television Tubes

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7.2.2 Storage Tubes

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7.3 Gas Tubes

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H. H. Wittenberg

7.3.1 Task Group on Methods of Test for TR and ATR Tubes

K. Garoff, *Chairman*
J. Schussele, *Secretary*

I. Birnbaum
W. R. Hayter
H. Heins
F. McCarthy
L. W. Roberts
R. Scudder

7.4 Camera Tubes, Phototubes, and Storage Tubes in Which Photo-emission is Essential

R. G. Stoudenheimer, *Chairman*

B. R. Linden
B. H. Vine
J. E. White

7.5 Test Methods for High-Vacuum Microwave Tubes

E. M. Boone, *Chairman*

J. H. Bryant
H. W. Cole
G. A. Espersen
L. M. Field
M. S. Glass
A. E. Harrison
M. E. Hines
R. A. LaPlante
R. R. Moats
C. R. Moster
M. Nowogrodzki
W. G. Shepherd

7.5.1 Task Group on Non-Operating Characteristics of Microwave Tubes

M. Nowogrodzki, *Chairman*

B. Arfin
E. M. Boone
M. S. Glass
R. C. Hergenrother
E. D. Reed

7.5.2 Task Group on Operating Measurements of Microwave Oscillator Tubes

R. R. Moats, *Chairman*

R. S. Briggs
T. P. Curtis
C. Dodd
G. I. Klein
B. D. Kumpfer
E. C. Okress
W. Teich

7.5.2 Consultants

G. E. Hackley
A. E. Harrison
O. C. Lundstrom
E. D. McArthur
J. S. Needle
W. G. Shepherd

7.5.3 Task Group on Operating Measurements of Microwave Amplifier Tubes

H. W. Cole, *Chairman*

J. Berlin
H. Hersh
R. G. E. Hutter
P. M. Lally

A. W. McEwan
C. R. Moster
R. W. Peter
S. E. Webber

Consultants

J. H. Bryant
M. Chodorow
C. C. Cutler
L. M. Field
A. E. Harrison
M. E. Hines
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L. D. Smullin
D. Watkins

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7.7 Solid State Devices

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C. L. Rouault
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A. G. Coblenz

7.7.2 Methods of Test for Transistors for Linear CW Transmission Service

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A. E. Anderson
A. Coblenz
L. J. Giacoletto
A. W. Lampe

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M. F. Lamonte
B. J. Rothlein
J. S. Schaffner
K. Uhler

7.7.4 Methods of Test for Transistors for Pulse Service

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J. C. Lozier
W. M. Pease
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P. Travers
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S. B. Williams
F. R. Zatlin

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V. Azgapetian, *Secretary*

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T. Flynn
J. C. Lozier
C. F. Rehberg
F. Zweig

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J. Weinberger

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J. M. Cage
J. Dalke
E. Mittelmann
W. Richter

10.2 Dielectric Measurements

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H. R. Meahl
J. H. Mennie
C. F. Spitzer

10.4 Methods of Measurement

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J. L. Dalke
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W. R. Thurston

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H. W. Kohler
E. R. Kretzmer
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N. Marchand
L. A. Meacham
D. Pollack

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J. H. Mulligan, Jr., *Vice-Chairman*

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J. L. Dalke
G. L. Fredendall
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G. B. Hoadley
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G. A. Morton
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A. P. G. Peterson
J. G. Reid, Jr.
R. Showers

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25.1 Basic Standards and Calibration Methods

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G. L. Davies
F. J. Gaffney

25.2 Dielectric Measurements

J. L. Dalke, *Chairman*

C. A. Bieling
F. A. Muller

25.3 Magnetic Measurements

C. D. Owens, *Chairman*

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G. S. Fay
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25.5 Video Frequency Measurements

G. L. Fredendall, *Chairman*

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W. R. Thurston

25.6 High Frequency Measurements

R. V. Lowman, *Chairman* Joint AIEE-IRE Committee on High Frequency Measurements
G. B. Hoadley, *Chairman*, IRE Subcommittee 25.6

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I. G. Easton
F. J. Gaffney
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B. M. Oliver
B. Parzen

25.8 Interference Measurements

R. Showers, *Chairman*

F. H. Dickson
H. E. Dinger
C. W. Frick
F. M. Greene
A. W. Sullivan

25.9 Measurement of Radio Activity

G. A. Morton, *Chairman*

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D. L. Collins
L. Costrell
D. L. Johnston
M. A. Schultz
W. W. Schultz

25.10 Oscillography

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P. S. Christaldi
M. E. Clark
D. P. Heacock
H. M. Joseph
G. R. Mezger (*alternate*)
N. E. Schick (*alternate*)
A. L. Stillwell
H. Vollum

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F. L. Verwiebe
G. F. C. Weedon
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25.14 Electronic Components

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J. W. Gruol
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A. E. Javitz
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24.7 Terrestrial Radio Noise

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National Research Council—Division of Engineering and Industrial Reserch: F. B. Llewellyn (7/1/54–6/30/57), 3 yrs.

American Association for the Advancement of Science: Professor J. C. Jensen (7/1/54–6/30–56), 2 yrs.

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AERONAUTICAL AND NAVIGATIONAL ELECTRONICS

Vol. ANE-1, No. 2, June, 1954

Report on the Dayton Conference—K. C. Black

The Long Quest—P. C. Sandretto (Editorial)

Radio Altimeter—M. Capelli

This paper discusses the fundamental principles underlying the operation of radio altimeters of the frequency modulation type. The discussion is related to work done by STC in development of an accurate short-range radio altimeter. Ideal and practical solutions of the problems involved are explained by means of a number of graphs and discussions of such points of interest as type of frequency sweep, instantaneous frequency measurement versus average rate counting, step effect and minimum range. The paper ends with a specification of operating parameters for a practical radio altimeter.

Flare-Out Unit AN/APN-71, an Aid to Aircraft Instrument Landing—D. M. Pasek and W. J. Shanahan

Navy's New Altimeter, the AN/APN-22—F. T. Wimberly and J. F. Lane, Jr.

This paper describes a frequency modulation type of radio altimeter developed for the Navy Department, Bureau of Aeronautics. It operates in the frequency band 4,200 to 4,400 megacycles and has a range of 0–10,000 feet over land and 0–20,000 feet over water. The accuracy is ±2 feet from 0 to 40 feet and ±5% from 40 feet to 20,000 feet. The equipment is designed to meet standard Navy specifications as to environmental conditions. Total weight is 30 pounds. Altitude is displayed on a single dial indicator, linearly calibrated over 120 degrees for the range 0–200 feet, and approximately logarithmically for the remaining 190 degrees.

In addition to the indicator, the equipment comprises a transmitter-receiver and a control amplifier. The former is flush mounted and contains the antennas, transmitting magnetron, and receiver crystals. The latter generates the basic FM sweep voltage, provides synch signals to operate the indicator, and furnishes the power for the complete system.

A feature of the system is a flag alarm warning which operates when the received signal-to-noise ratio is too low for reliable indications. The system will operate in banks up to 50° and dives and climbs of 60°.

Nonquantized Frequency-Modulated Altimeter—H. P. Kalmus, J. C. Cacheris and H. A. Dropkin

A new principle that makes possible nonquantized altitude information so that the accurate range of altimeters can be extended to include low-level operation is described. A frequency shifter is inserted between the local oscillator and the mixer in order to remove the inherent error in conventional altimeters that makes them objectionable for low-altitude operation. The authors show how the characteristic of the amplifier can be shaped differently and point out that the strength of the return signal and the inherent amplitude modulation determine the most suitable characteristic for any particular operating condition.

ANTENNAS AND PROPAGATION

Vol. AP-2, No. 4, October, 1954

News and Views

Analysis of Helical Transmission Lines by Means of the Complete Circuit Equations—V. J. Fowler

A set of integro-differential equations, called the "complete circuit equations," are derived from Maxwell's equations and applied to the solution of the parallel-wire transmission line, the double-helix transmission line, and the single helix, or helical waveguide. These equations take into account the effects of inductance and capacitance distribution, retardation, and outward radiation. A generalization of earlier concepts of distributed inductance and elastance (or inverse capacitance) is manifest in the solution of the helical line, where these quantities become functions of the phase coefficient or wavelength of propagation and are Fourier transforms of certain closed-form distribution functions. In general, phase velocity is a complicated implicit function of frequency, but under a hypothesis of "mode segregation on the basis of wavelength," the phase velocity and frequency can be obtained parametrically in terms of a third variable, called the phase parameter. Using this hypothesis, plots of phase velocity and characteristic impedance versus frequency were obtained for the double helix and the helical waveguide.

Radiation From a Vertical Dipole Over a Stratified Ground—J. R. Wait and W. A. Fraser

Further results are given for the problem of a vertical electrical dipole situated over a horizontally stratified conductor. It is pointed out that under certain conditions, the surface-wave field intensity for a stratified conducting ground is greater than the corresponding case for a perfectly conducting ground. Numerical values for the attenuation factor are also given.

A Waveguide Array for Solar Noise Studies—H. Gruenberg

A description is given of a 150-foot slotted waveguide array which was built for solar noise studies at wavelengths near 10.3 centimeters. Some of the problems associated with the design of such an array are discussed.

Dielectric Sheet Radiators—F. E. Butterfield

A class of electromagnetic radiators is described which employs the principle of waveguiding along a flat surface by means of dielectric coating. Radiation occurs as the result of nonuniformities in the guiding system. The efficiency of the feeding arrangements and diffraction over the edge of the flat surface are factors in the side-lobe level observed. A unit is described which has been used aboard missiles.

A series of tests are reported which illustrate some of the characteristics of the radiators. Data for comparison with corrugated surface antennas is given. Problems for further investigation are listed.

An experimental unit is described with a gain of 25 db and an efficiency compared to conventional aperture radiators of about 60 per cent. This unit has linear dimensions comparable to a horn producing the same beamwidth.

Evaluation of Errors in an Eight-Element Adcock Antenna—J. R. Wait and W. A. Pope

An analysis is given for the response of an eight-element direction finding antenna to a localized radio-frequency source. The error between the indicated and true bearing is evaluated and illustrated by graphs. It is also shown that the additional error, introduced by bringing the source into proximity of the antenna system is negligible if the antenna-source distance is greater than 5λ. This is an important consideration in the calibration of the system.

An Exact Step-Up Impedance Ratio Chart of a Folded Antenna—Y. Mushiake

ELECTRON DEVICES

Vol. ED-1, No. 3, August, 1954

Traveling Wave Tube Characteristics for Finite Values of C—C. K. Birdsall and G. R. Brewer

Using the determinantal equation for the propagation constants of the waves in a traveling wave tube, values of the Pierce wave parameter $\delta = x + jy$ have been found for finite values of the gain parameter C (up to $C = .5$) and over a wide range of values of the space charge parameter QC (up to $QC = 2$). Using the values of the wave parameters, graphs were obtained of several derived quantities useful in traveling wave tube design. The following are given as functions of C and QC: (a) the initial loss factor (A_1) and the space charge loss factor (A_2) where $A_1 + A_2 = A$ of Pierce, (b) maximum values of the gain per wavelength parameter x_1, (c) maximum values of the gain per meter parameter $x_1/(1+Cb)$, (d) stream velocity parameter b at maximum gain per meter, (e) and (f) stream and growing wave phase velocity at maximum gain per meter as a function of $\sqrt{4QC}/(1-\sqrt{4QC^3}) \cong \omega/(\omega_q C)$.

Transistor Metrology—D. A. Alsberg

The general transistor measurement problem is surveyed considering the transistor designer, manufacturer and user. Several possible sets of transistor parameters in common use are discussed, stressing consistency between problems of measurement and the nature of transistors. For general small signal characterization, the hybrid parameters are proposed as the preferred set. For very high frequencies, finite termination, insertion type parameters are proposed.

The relative merits of point by point and sweep type measurements are discussed. Problems in both recording and oscilloscopic sweep measurements are considered. Techniques originally developed for rapid, precise measurement of long distance broad band transmission facilities may be adapted to precise transistor sweep instrumentation. This permits practical specification of transistors in terms of "regional specification" rather than the more conventional specification limits at discrete operating points.

A Developmental Voltage-Tunable Microwave Magnetron—J. S. Needle

The paper concerns the results of design and performance of a low power coaxially interdigital magnetron with external coupled cavity. Special attention is given to the problem of wide-range tunability, in the microwave frequency spectrum, using either mechanical or voltage-tuning methods.

The basic structure of the magnetron consists of a short, hermetically-sealed section of coaxial transmission line which contains six equally spaced radial vanes extending from the inner wall of the outer coaxial cylinder into six longitudinal slots in the inner coaxial cylinder. The cathode is symmetrically located within the inner coaxial cylinder at the position of the multi-anode structure. This sealed off multi-anode system is used to excite a TEM mode in an external coaxial circuit.

Experimental results on electronically and mechanically tunable operation are reported, with particular emphasis on voltage tuning.

Measurement of Klystron Amplifier Parameters—C. M. Beronda

This paper describes certain basic design measurements made on a high power klystron amplifier developed by Sperry under an Air Force (Rome Air Development Center) contract. This work was done by the Electronic Tube Department at Sperry during 1948–1950.

The klystron has a power output in the megawatt region and a high level gain of 38 db. The electron gun perveance is about 1×10^{-6} ampvolts$^{-3/2}$. About 95% of the electron gun current reaches the collector under static conditions while about 90% goes through under operating conditions. A plate efficiency of about 43% has been measured with up to 4 megawatts of peak power output.

Rather extensive engineering measurements were made on the klystron and these measurements check rather closely with the original design parameters. Among the tests made on the tube were the following measurements which are described:

1. The optimum drive 1st and 2nd resonator voltage ratios.
2. Beam loading measurements.
3. Longitudinal debunching measurements.
4. Large signal measurements.

Calculations of Wave Propagation on a Helix in the Attenuation Region—S. E. Webber

Calculations are made of the wave velocity and attenuation of a wave propagating on a helical current sheath in a housing. An attenuator of variable conductivity is assumed to be coincident with the helical current sheath. Results show that the wave velocity increases with increasing conductivity while the attenuation goes through a maximum value for a particular value of conductivity. These calculations are compared with measured values.

Transistors and Their Applications (A Bibliography, 1948–1953)—A. R. Krull

This bibliography consists of about 1,000 items, 1948–1953, arranged by year and by author, with an appended supplement. No annotations or abstracts are given.

ELECTRONIC COMPUTERS

Vol. EC-3, No. 3, September, 1954

The Editor

A Permanent High Speed Store for Use With Digital Computers—R. D. Ryan

A new type of high speed store is proposed for an electronic digital computer using interpretive program techniques. The store is based on the flying spot technique used in television signal generation. The information in the store may be read rapidly but is nonerasable. This store has the advantages of high storage density, good reliability and nonvolatility of information.

Application of Boolean Algebra to Switching Circuit Design and to Error Detection—D. E. Muller

A solution is sought to the general problem of simplifying switching circuits that have more than one output. The mathematical treatment of the problem applies only to circuits that may be represented by "polynomials" in Boolean algebra. It is shown that certain parts of the multiple output problem for such circuits may be reduced to a single output problem whose inputs are equal in number to the sum of the numbers of inputs and outputs in the original problem. A particularly simple reduction may be effected in the case of two outputs.

Various techniques are described for simplifying Boolean expressions, called "+ polynomials," in which the operation "exclusive or" appears between terms. The methods described are particularly suitable for use with an automatic computer, and have been tested on the Illiac.

An unexpected metric relationship is shown to exist between the members of certain classes of "+ polynomials" called "nets." This relationship may be used for constructing error-detecting codes, provided the number of bits in the code is a power of two.

An Algebraic Theory for Use in Digital Computer Design—E. C. Nelson

An algebraic theory of the logical operation of digital computers is developed. This theory takes into account the dynamic (time) behavior of computer processes. The computer signals and computer elements are described. Their properties which are pertinent to the logical operation of digital computers are abstracted and formulated in mathematical terms. The signals are represented by algebraic symbols, and the way they are transformed by the elements of the computer is represented in terms of algebraic operations and functions. This computer algebra is based on Boolean algebra. Time is treated as a discrete variable and a method of taking into account the time relationships in the computer process is developed. Specific components, such as gates, flip-flops, and magnetic drums, are analyzed, and an algebraic description of their operation is obtained.

An Improved Reading System for Magnetically Recorded Digital Data—Samuel Lubkin

In magnetic recording of pulses, whether on drum or tape, the resulting flux pattern is affected by proximity of adjacent pulses. The best defined region is that adjacent to the maximum. In reading, the signal is the derivative of the flux. In the best defined region, this is close to a straight line passing through zero when the flux is maximum. The slope of the curve at the zero changes sign with change of pulse polarity. A new method of reading is described which examines the signal from the head for such transitions from positive to negative or reverse as indications that a positive or negative pulse had been recorded. This is done by gating the inverted signal with the delayed signal for positive pulse reading and the inverse of this for reading negative pulses. Besides providing sharply defined outputs, this method permits reading both positive and negative pulses from a single channel without interference or ambiguity. Examples are given for using this facility for checking purposes and for storage of two types of data in a common channel.

A Digital Voltage Encoder—J. R. Zweizig

A two-channel voltage encoder having a sampling rate of 40 numbers/second in each channel has been designed for use in a data reduction system. The data are recorded on single-channel magnetic tape in the form of 10-digit binary numbers with an accuracy of ±0.1 per cent. The range of input voltage is 0 to 1 volt and may be lowered to 0 to 10 mv through the use of chopper amplifiers. This lower range of input voltage covers the voltages generated by analog transducers commonly employed to measure temperatures, pressures, flow rates, and thrusts.

A New Method of Generating Functions—Lazarus G. Polimerou

As a result of a pressing need for function generators, a new method of function generation has been developed. The underlying principle of this function generator is the application of ordinary pulse techniques in such a way as to produce a function. The simplicity of design, the high accuracy attainable, the simple type of construction are the outstanding features of this general-purpose function generator.

In order to compare this new type of function generator with those presently being used, three important types are discussed. These generators are of the general-purpose, electric and photoelectric types; other comparable electromechanical types are excluded.

A Function Generator for the Solution of Engineering Design Problems—C. J. Savant and R. C. Howard

The solution of nonlinear engineering design problems demonstrates the need for a special function generator. The generator described in this paper satisfies this need. The basic components of the unit are discussed and the forms of functions which can be generated are shown. Accuracy is estimated by comparison of an oscillogram with the calculated curves. It is concluded from tests on the system that the function generator is a valuable aid in the handling of nonlinear design problems.

News

Contributors

Review Section

Abstracts and References

Compiled by the Radio Research Organization of the Department of Scientific and Industrial Research, London, England, and Published by Arrangement with that Department and the *Wireless Engineer*, London, England

NOTE: The Institute of Radio Engineers does not have available copies of the publications mentioned in these pages, nor does it have reprints of the articles abstracted. Correspondence regarding these articles and requests for their procurement should be addressed to the individual publications, not to the IRE.

Acoustics and Audio Frequencies....... 1589
Antennas and Transmission Lines...... 1589
Automatic Computers................. 1590
Circuits and Circuit Elements.......... 1590
General Physics..................... 1592
Geophysical and Extraterrestrial Phenomena........................ 1593
Location and Aids to Navigation....... 1595
Materials and Subsidiary Techniques... 1595
Mathematics........................ 1597
Measurements and Test Gear......... 1597
Other Applications of Radio and Electronics........................ 1597
Propagation of Waves................ 1598
Reception.......................... 1599
Stations and Communications Systems.. 1599
Subsidiary Apparatus................ 1600
Television and Phototelegraphy........ 1600
Transmission....................... 1601
Tubes and Thermionics............... 1601

The Index to the Abstracts and References published in the PROC. I.R.E. from February 1953 through January 1954 is published by *Wireless Engineer* and included in the March 1954 issue of that journal. Copies of this issue may be purchased for $1 (including postage) from the Institute of Radio Engineers, 1 East 79th Street, New York 21 N. Y. As supplies are limited, the publishers ask us to stress the need for early application for copies. Included with the Index is a selected list of journals scanned for abstracting with publishers' addresses.

The number in heavy type at the upper left of each Abstract is its Universal Decimal Classification number and is not to be confused with the Decimal Classification used by the United States National Bureau of Standards. The number in heavy type at the top right is the serial number of the Abstract. DC numbers marked with a dagger (†) must be regarded as provisional.

A new section has been introduced covering the general technique of electromagnetic waves, oscillations, and pulses (i.e. transmission lines and circuits), as distinct from specific applications to telecommunications. The new section is numbered 621.37, with subdivisions. Full details of the new classification, and of the numbers which become obsolete as a result of its introduction, are given in PE Note 535, obtainable from The International Federation for Documentation, Willem Witsenplein 6, The Hague, Netherlands, or from The British Standards Institution, 2 Park Street, London, W.1., England.

Section 621.396.67, dealing with Antennas, has been modified and expanded; details of the new classifications are given in PE Note 519.

Section 621.396.96, with subdivisions, has been introduced to cover Radar; details of the new classifications are given in PE Note 518.

New Subject Section

A section headed Automatic Computers has been introduced.

ACOUSTICS AND AUDIO FREQUENCIES

534.232:534.321.9 2550

Devices for generating Ultrasonic Waves in Air, using Vibrating Cylinders, Pistons, Spheres and Cubes—V. Gavreau and M. Miane. (*Acustica*, vol. 4, pp. 387–395; 1954. In French.) Electrodynamic types of ultrasonic generators are particularly suitable for laboratory acoustic measurements. Focused arrangements and regenerative systems for high frequencies (77 kc) are described. Visual methods for studying the vibrations of the emitting surfaces are discussed. Application to the absolute calibration of microphones at ultrasonic frequencies is indicated.

534.44 2551

Investigation of Nonperiodic [noise] Processes by means of Autocorrelation and Fourier Analysis—M. L. Exner. (*Acustica*, vol. 4, pp. 365–379; 1954. In German.) The two methods were compared by analyzing filtered noises and hisses. For frequency and bandwidth determinations the Fourier-analysis method is preferred, but for energy-distribution determinations the autocorrelation method is better.

534.614 2552

The Effect of Slip on Sound Propagation—D. E. Weston. [*Proc. Phys. Soc.* vol. 67, p. 265; March 1 1954.] Comment on 615 of March (Caro and Martin).

534.64 2553

Acoustic Impedance Measurement using a Resonance Method—M.A. Ferrero and G. G. Sacerdote. (*Acustica*, 1954, vol. 4, pp. 359–364; 1954.) See 2545 of 1953.

534.64 2554

Determination of High Mechanical Input Impedances of Solid Bodies in the Frequency Range 50–3000 c/s—W. Elling. (*Acustica*, vol. 4, pp. 396–402; 1954. In German.) An electromechanical vibrometer is described and measurements on walls, ceilings and Al plates are reported. The results show that the reactive part of the impedance of ceilings and walls is an inertance at low frequencies and a stiffness at high frequencies; the change-over occurs at about 150–300 cps, where the resistive part of the impedance is a maximum.

534.78 2555

The Experimental Study of Speech—D. B. Fry. [*Nature* (*London*) vol. 173, pp 844–846; May 8, 1954] A survey of recent work on the analysis and synthesis of speech.

534.845 2556

Improvement of the Effectiveness of Acoustic Panels by use of Resonators—G. Kurtze. (*Tech. Mitt. schweiz. Telegr.-TelephVerw.*, vol. 32, pp. 81–87; March 1, 1954.) The absorptive effect of the panel material is supplemented, particularly at low frequencies, by providing circular holes or slits, so as to form resonators. Empirical formulas are used for the calculations. Absorption/frequency characteristics and studio reverberation-time/frequency curves illustrate the improvement attained.

534.846.6 2557

Polycylindrical Systems and the Diffusion of Sound Waves—R. Lamoral. (*Onde élect.*, vol. 34, pp. 308–309; March, 1954.) A study was made of the suppression of echoes in rooms by means of polycylindrical surfaces, using 1:50 reduced-scale models and frequencies between 5 and 30 kc. The results are presented in a family of curves for different distances between source and reflecting surface. Echo suppression improves as this distance increases, and at all distances is best for wavelengths between 0.5 d and d (l = cylinder chord).

621.395.6 2558

Developments in Sound Reproduction—(*Wireless World*, vol. 60, pp. 313–317; July, 1954.) Review of products shown at exhibitions held by the Radio and Electronic Component Manufacturers' Federation, the Association of Public Address Engineers, and the British Sound Recording Association.

621.395.61:621.391.5 2559

The "Vagabond" Wireless Microphone System—Phinney. (*See* 2765.)

621.395.625 2560

Congress on Sound Recording—Paris 1954—H. J. Houlgate. (*Wireless World*, vol. 60, pp. 348–350; July, 1954.) About 60 papers were presented and discussed at the "International Conference on Sound Recording Processes and their Extension to the Recording of Information," held in April 1954. Points from the discussions and items of interest in the associated exhibition are noted. The full text of many of the papers and abstracts of others are given in *Onde élect.*, vol. 34, pp. 193–294; March, 1954.

621.395.625.3 2561

Synchronized Magnetic Tape Recording—R. H. Ranger. (*Trans. AIEE*, Part I, *Communication and Electronics*, vol. 72, pp. 581–586; 1953.) The longitudinally-magnetized sound track on a 1/4-inch tape for use with film is synchronized by recording on the same tape a control track consisting of transverse-magnetization signals derived from the ac power driving both the tape and the camera.

ANTENNAS AND TRANSMISSION LINES

621.372 2562

Coupled Wave Theory and Waveguide Applications—S. E. Miller. (*Bell Sys. Tech. Jour.*, vol. 33, pp. 661–719; May, 1954.) Theory applicable to coupled transmission lines, including waveguides and helical lines, is developed. For loose coupling, when very little power is transferred, the length of the coupling region can be reduced by tapering the coupling. For tight coupling distributed uniformly in the direction of propagation, energy is exchanged periodically provided that the attenuation and phase constants of the two lines are equal or nearly so. Experiments have confirmed the theory. Applications include pure-mode couplers in multimode systems, and wave filters.

621.372.8:[537.562+621.318.134 2563

Topics in Guided-Wave Propagation through Gyromagnetic Media: Part 1—The Completely Filled Cylindrical Guide—H. Suhl and L. R. Walker. (*Bell Sys. Tech. Jour.*, vol.

33, pp. 579–659; May, 1954.) "The characteristic equation for the propagation constants of waves in a filled circular guide of arbitrary radius is written in terms of magnetizing field and a carrier density, which are shown essentially to determine the dielectric and permeability tensors for a gas discharge plasma and for a ferrite. The complex structure of the spectrum of propagation constants and its dependence upon radius and the two parameters are analyzed by a semigraphical method, supplemented by exact formulas in special regions. Thus the course of individual modes may be charted with fair accuracy."

621.372.8:538.614 2564

Gyrotropic Waveguide—M. A. Gintsburg. (*Compt. Rend. Acad. Sci.* (*URSS*), vol. 95, pp. 489–492; March 21, 1954. In Russian.) The theory is developed of em propagation in a bounded anisotropic medium, in particular in waveguides which are partly or completely filled with a medium with μ_{ik} and ϵ_{ik} varying with direction. Such "gyrotropic" (rotation-producing) media can be obtained by using the Faraday or Kerr effects or similar phenomena. The final set of equations derived indicates a method of determining the components of the nonsymmetric tensors representing μ_{ik} and ϵ_{ik} by means of a resonator. See also 2710 of 1952 (Suhl and Walker).

621.396.67 2565

An Experimental Study of a Microwave Periscope—J. Drexler. (Proc. I.R.E., vol. 42, p. 1022; June, 1954.) Experiments were made to test the validity of the theoretical result obtained by Jakes (1243 of 1953) that it is possible to obtain more power from a combination of antenna and reflector than from the same antenna located at the reflector position. For one particular case the gain of the combination was 2.3 db above that of the antenna alone.

621.396.67:621.397.6 2566

Cosecant Antenna aids U.H.F.-TV Coverage—J. E. Martin and J. Ruze. (*Electronics*, vol. 27, pp. 138–142; June, 1954.) An antenna designed to give uniform field strength over the service area comprises four vertical tubes of structural steel, of outer diameter 4 inches, arranged with their cross sections in a square, and their axes 5 inches apart. These tubes constitute the outer conductor of a coaxial system. Rings of slot radiators with centers spaced one wavelength apart are formed between steel welding members. Details are given of feed arrangements and radiation patterns obtained. Coverage was tested by operation at a television station on 790 mc.

621.396.67.012.12:517.864 2567

The Maximal Directivity Coefficient of Linear and Plane Aerials—L. D. Bakhrakh. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 95, pp. 45–48; March 1, 1954. In Russian.] The radiation-pattern function is given in terms of a series of Mathieu functions of antenna length, l, in wavelengths, λ, and of direction angle, of orders up to the nth (see 1675 of June). The condition for the directivity coefficient to be a maximum in a given direction, for a given value of n, and expressions for the resultant field and antenna losses are derived. Results of calculations for $l/\lambda = 10.58$ are shown graphically.

621.396.676:621.396.93 2568

U.H.F. Omnidirectional Antenna Systems for Large Aircraft—W. Sichak and J. J. Nail. (*Trans. I.R.E.*, vol. AP-2, pp. 6–15; Jan., 1954.) Experiments have been made on models to investigate the siting of antennas on different types of aircraft. Analysis of results is based on the probability of attaining free-space-dipole range for all azimuth angles and ±30-degree elevation. The operation of dual antenna systems for navigation and traffic-control equipment is discussed. Direct parallel feed of dual antennas gives an interference region where individual radiation patterns overlap; in this region, a radar beacon may operate satisfactorily but the performance of dme is uncertain. Delay-line, switching and phase-shift systems of operation are considered. The simplest common-antenna arrangement is the operation of nose and tail antennas in parallel, with hybrid multiplexing.

621.396.677.3 2569

A Method for Calculating the Current Distribution of Tschebyscheff Arrays—D. Barbiere. (Proc. I.R.E., vol. 42, p. 1021; June, 1954.) Discussion on p. 1211 of 1952.

621.396.677.43 2570

Rhombic Antennas for Transmitting Stations—L. Long. (*Brown Boveri Rev.*, vol. 40, pp. 407–416; October, 1953.) The radiation diagram of a rhombic antenna is analyzed, taking account of the attenuation due to radiation. Values of maximum gain attainable with a single rhombic of characteristic impedance 600Ω are noted and the design and application of single and double rhombic antennas for point-to-point transmission and for transmission to a specific zone are considered.

621.396.677.43.011.21 2571

Mutual Impedance of Stacked Rhombic Antennas—J. G. Chaney. (*Trans. I.R.E.*, vol. AP-2, p. 39; January, 1954.) A formula derived earlier (3199 of 1953) is integrated for the case of two identical rhombic antennas; the resultant expression involves associated sine and cosine integral functions.

621.396.677.71 2572

Mutual Coupling Considerations in Linear-Slot Array Design—M. J. Ehrlich and J. Short. (Proc. I.R.E., vol. 42, pp. 956–961; June, 1954.) An experimental study was made of the external coupling between slots in adjacent resonant waveguides on a finite ground plane. Measurements were made of the field strength in one of the guides on exciting the other guide, and of the change in input admittance of the driven slot due to the presence of the parasitic slot, with matched termination of the parasitic waveguide. The results indicate that the slot-coupling effect can generally be neglected in the design of linear-slot arrays.

621.396.677.83:621.396.96 2573

Torque Requirements of a Radar Antenna—M. Mark. (*Elec. Eng.*, vol. 73, pp. 262–264; March, 1954.) Results are given of wind-tunnel tests on an experimental radar antenna. The torque required to rotate the antenna varied with azimuth position, elevation angle, pivot location, speed of rotation, and wind velocity. The combined results are evaluated on the basis of dimensionless parameters.

621.396.677.85 2574

The Electromagnetic Field near a Dielectric Lens—D. C. Hogg. (*Jour. Appl. Phys.*, vol. 25, p. 542; April, 1954.) The field components along the electric and magnetic diameters differ considerably near the lens surface. The spacing between successive maxima of field intensity along the magnetic diameter near the optic axis is of the order of the wavelength in the dielectric. Virtual sources along the aperture edge, as suggested by Andrews (3141 of 1950), can account for these observations.

AUTOMATIC COMPUTERS

681.142 2575

Solutions of Boundary-Value Problems on Automatic Computing Equipment—F. M. Verzuh. (*Trans. AIEE*, Part I, *Communication and Electronics*, vol. 72, pp. 813–821; 1953.) Report of an investigation in which five types of digital computer and one differential analyzer were compared as regards their suitability for solving boundary-value problems.

681.142 2576

A Progressive Code Digital Quantizer—F. Raasch. (*Trans. AIEE*, Part I, *Communication and Electronics*, vol. 72, pp. 567–571; 1953.) Description of an analog-to-digital converter which receives information from a graph reader and gives a corresponding numerical indication. Several circuits are shown, based on the principle of comparing the analog voltage with the sum of a finite geometric series.

681.142 2577

The Manchester University High-Speed Digital Computer—D. B. G. Edwards. (*Jour. Brit. IRE*, vol. 14, pp. 269–278; June, 1954.) See 978 (Kilburn et al.) and 979 (Pollard and Lonsdale) of April.

681.142 2578

A Mercury Delay-Line Memory Unit—R. D. Ryan [*Proc. IRE*, (*Australia*), vol. 15, pp. 89–95; April, 1954.] The storage unit in the CSIRO digital computer [1050 of 1953 (Beard and Pearcey)] is described; a method is discussed of interspacing the pulses of one loop between those of another loop, thus doubling the storage capacity.

681.142 2579

A Review of Magnetic and Ferroelectric Computing Components—V. L. Newhouse. (*Elec. Eng.*, vol. 26, pp. 192–199; May, 1954.) High-speed digital storage devices may be classified as delay-line or random-access types, with subdivisions in each case for regenerative and nonregenerative devices. The more important magnetic and ferroelectric storage units are critically discussed and possible future lines of development indicated. 26 references.

681.142 2580

An Arithmetic Unit for Automatic Digital Computers—J. R. Stock. (*Z. angew. Phys.*, vol. 5, pp. 168–172; March 15, 1954. In English.) Discussion of the requirements for the arithmetic unit of a computer with magnetic-drum store, with particular reference to the computer for the Swiss Federal Institute of Technology. Operation with fixed or floating decimal point is provided for.

681.142 2581

Magnetic Switching Circuits—R. C. Minnick. (*Jour. Appl. Phys.*, vol. 25, pp. 479–485; April, 1954.) Magnetic circuits to produce the logical operations of "and" and "or" are described, which may be combined to produce any switching function of any number of input binary variables. The only components used are magnetic cores, wire, and resistors. There is no extremely sensitive parameter in these circuits which must be adjusted carefully in order to ensure proper operation.

681.142:512 2582

A Simplified Solution and New Application of an Analyser of Algebraic Polynomials—L. Lukaszewicz. (*Bull. Acad. Polon. Sci.*, vol. 1, pp. 103–107; 1953. In English.) A description is given of an analyzer circuit in which complex numbers are represented by sinusoidal voltages of frequency ∼500 cps, with amplitude corresponding to the modulus and phase corresponding to the argument of the number.

681.142:621.385.832 2583

An Electron-Beam Tube for Analog Multiplication—E. J. Angelo, Jr. (*Rev. Sci. Instr.*, vol. 25, pp. 280–284; March, 1954.) An electron beam of large circular cross section and uniform current density is projected through an es deflection systen on to a metallic target made up of four quadrants insulated from one another. The currents collected by one diagonal pair of quadrants are arranged to oppose those collected by the other pair, thus the algebraic sum of the currents in the quadrants is proportional to the product of the deflection voltages. The design and performance of the system is discussed.

CIRCUITS AND CIRCUIT ELEMENTS

621.314.2 2584

Prevention of Ionization in Small Power Transformers—L. Medina. (*Proc. IRE*, (*Aus-

tralia), vol. 15, pp. 114–115; May, 1954.) The ionization-onset voltage is raised, for a given total thickness of insulation, by inserting foils in the insulation so as to split it into n sections with $1/n$ of the total voltage across each section. Design details and test results for a typical transformer are given.

621.316.8+621.319.4 2585

Some Characteristics and Limitations of Capacitor and Resistor Components—L. Podolsky and J. K. Sprague. (*Trans. I.R.E.*, no. PGCP-1, pp. 33–46; March, 1954.) Six types of film-dielectric capacitors are examined with respect to their performance. Specifications for a new and reliable type of metallized paper capacitor, at the pilot-plant development stage, are given. A satisfactory method for power rating of precision bobbin resistors is explained.

621.318.134:621.318.5 2586

Stressed Ferrites having Rectangular Hysteresis Loops—J H. Williams, R. C. Sherwood, M. Goertz and F. J. Schnettler. (*Trans. AIEE*, Part I, *Communication and Electronics*, vol. 72, pp. 531–537; 1953.) To obtain the rectangular hysteresis loop required in ferrite cores used in switching and storage devices, the cores are stressed by encasing them in plastics which shrink during polymerization.

621.318.4 2587

Temperature Fields in Electrical Coils: Numerical Solutions—P. J. Schneider. (*Trans. AIEE*, Part I, *Communication and Electronics*, vol. 72, pp. 768–771; 1953.) The computation of the steady temperature fields in coils by the relaxation method is described and illustrated.

621.318.435 2588

Saturable Reactors with Inductive D.C. Load: Part 2—Transient Response—H. F. Storm. (*Trans. AIEE*, Part I, *Communication and Electronics*, vol. 72, pp. 182–192; 1953.) Part 1: 3527 of 1953.)

621.318.57 2589

A Graphical Method for Flip-Flop Design—R. F. Johnston and A. G. Ratz. (*Trans. AIEE* Part I, *Communication and Electronics*, vol. 72, pp. 52–56; 1953.)

621.318.57 2590

Switching in Bistable Circuits—R. S. Mackay. (*Jour. Appl. Phys.*, vol. 25, pp. 424–429; April, 1954.) All triggering processes seem capable of description in terms of a curve with a region of negative slope, points on this curve representing unstable states, while points on a positive slope represent stable states. A negative-resistance circuit in which the voltage is a multiple-valued function of current was constructed so that details of the triggering process could be observed on an oscilloscope screen. By suitable periodic sampling a slow-motion representation of the transition between two states is obtained.

621.318.57 2591

An Electronic Random Selector—R. W. Walker. (*Jour. Brit. IRE*, vol. 14, pp. 262–268; June, 1954.) Criteria for recognition of random series are stated and methods for producing such series are discussed. A description is given of an arrangement in which pulses are applied to a circuit which is gated periodically; any pulse passing through the gate triggers a multivibrator which is in turn coupled to a pair of mechanical registers. Two such circuits make up the complete selector enabling four mechanical registers to operate. Use of the selector in psychical research is mentioned.

621.318.57:621.314.7 2592

The Transient Response of Transistor Switching Circuits—I. L. Lebow and R. H. Baker. (Proc. I.R.E., vol. 42, pp. 938–943; June, 1954.) Transient response of the point-contact transistor in a grounded-base circuit is considered, using a linearized equivalent circuit. The required trigger voltages for both directions of switching are found experimentally to be greater than suggested by the static characteristic, and to depend on trigger pulse width to a different extent for the two directions. To fit these findings it is assumed that cut-off frequency is high when the transistor is switched off and low when it is switched on. The one-shot multivibrator is discussed in the light of these assumptions.

621.318.57:621.387 2593

A Cold-Cathode Scaling Unit—C. D. Florida and R. Williamson. (*Elec. Eng.*, vol. 26, pp. 186–191; May, 1954.) The five-decade scaler uses dekatrons, interstage coupling being by cold-cathode gas-filled trigger triodes. The resolving time is ~500 μs, which makes it suitable for application in conjunction with GM counter tubes. Operation under varying climatic conditions has been satisfactory.

621.319.4:621.315.612.4 2594

The Effect of Minor Constituents in High Dielectric Constant Titanate Capacitors—W. W. Coffeen. (*Trans. AIEE*, Part I, *Communication and Electronics*, vol. 72, pp. 704–709; 1953.) Additives used to modify the dielectric properties of $BaTiO_3$ are classified into two groups, those which alter the position of the Curie peak and those which lower its height. $BaSnO_3$ and $PbSnO_3$ are typical of the first group, while $MgSnO_3$, $Bi_3(SnO_3)_3$, $MgTiO_3$, and $CaTiO_3$, are typical of the second. The effect of composition on nonlinearity and aging is discussed. See also 191 and 192 of January.

621.319.4.001.4 2595

Breakdown and Leakage Resistance Investigation of Metallized Paper Capacitors—J. Burnham. (*Trans. I.R.E.*, no. PGCP-1, 3–17; March, 1954.) The basis of comparison between conventional paper and metallized-paper capacitors is discussed. Experiments on units with polyester impregnants suggest that, if self-healing is of primary importance, a pure aliphatic hydrocarbon type of impregnant will give best results.

621.372:517.948 2596

Applications of Integral Equations to the Solution of Nonlinear Electric Circuit Problems—L. A. Pipes. (*Trans. AIEE*, Part I, *Communication and Electronics*, vol. 72, pp. 445–450; 1953.) The technique described gives practical results in many cases where the approximate solution of the integral equation may be effected by means of Laplace transforms.

621.372.52+621.375.4 2597

Junction Transistor Circuits—(*Radio & Electronics* (*Wellington, N.Z.*), vol. 9, pp. 18–19, 23; March 1, 1954.) Application of the CK722 *p-n-p* junction-type transistor in single-stage and cascaded amplifiers and in lf and rf oscillators, is described and illustrated by eight circuit diagrams. The transistor collector characteristics, and operating data and typical circuit component values are given.

621.372.55 2598

Spectrum or Waveform Equalization?—D. A. Bell. (*Wireless Eng.*, vol. 31, pp. 171–174; July, 1954.) The method of equalization in which signals obtained by differentiation or integration are added to the main signal [1936 of 1953 (Gouriet)] is compared with the commoner method using filters. Empirical design is simpler with the former method, but the need for additional amplification at high frequencies is not avoided. The different requirements of communication and servo systems are indicated.

621.372.8:538.614 2599

New Effect caused by Gyromagnetic Phenomena—K. M. Polivanov. (*Compt. Rend. Acad. Sci.* (*URSS*), vol. 95, pp. 501–503; March 21, 1954. In Russian.) Theory is given of a device similar to that described by Kales et al. (2040 of July).

621.373 2600

Effects of Harmonics on the Frequency of Oscillation as well as on the Asymmetry of the Resonance Curves—A. E. Mostafa. (*Trans. AIEE*, Part I, *Communication and Electronics*, vol. 72, pp. 309–314; 1953.) The nonlinear equations representing the free and synchronized oscillations are transformed into linear equations with periodically varying coefficients, and complete solutions are obtained. The relation between frequency variation and harmonic content is derived. Experimental results supporting the theory are reported.

621.373 2601

The First Order Behavior of Separable Oscillators—D. C. dePackh. (*Trans. AIEE*, Part I, *Communication and Electronics*, vol. 72, pp. 450–455; 1953.) Analysis is given for oscillators whose linear and nonlinear elements are separable in the first order. Theoretical and experimental results indicate that, apart from adjustments of the linear elements, the major factor for obtaining frequency stability and smooth control is the suppression of the even-power terms in the amplitude characteristic.

621.373:621.396.822:530.145 2602

Quantum Theory of a Damped Electrical Oscillator and Noise: Part 2—The Radiation Resistance—J. Weber. (*Phys. Rev.*, vol. 94, pp. 211–215; April 15, 1954.) The results obtained in Part 1 (3243 of 1953) are extended to include damping and noise due to a radiation resistance and expressions for the mean square noise voltage and available noise power are derived which are shown to be the same as those previously obtained. They are also extended to apply to a single mode of a cavity resonator.

621.373:621.396.822:530.145 2603

Vacuum Fluctuation Noise—J. Weber. (*Phys. Rev.*, vol. 94, pp. 215–217; April 15, 1954.) The possibility of observing vacuum fluctuations by measuring the resultant electron stream noise when an electron stream interacts with a damped oscillator is discussed. An electron stream is shown to provide a means of precise measurement of the mean square noise emf for certain modes. The circuit must be at low temperature.

621.373.421+621.375.23].001.2 2604

Block-Diagram Solutions for Vacuum-Tube Circuits—T. M. Stout. (*Trans. AIEE*, Part I, *Communication and Electronics*, vol. 72, pp. 561–567; 1953.) Full paper. See 1346 of May.

621.373.421.13.029.65 2605

Millimeter Waves from Harmonic Generators—G. M. Johnson, D. M. Slager, and D. D. King. (*Rev. Sci. Instr.*, vol. 25, pp. 213–217; March, 1954.) Fundamental and harmonic powers obtainable at wavelengths down to 1.9 mm using a Si-crystal frequency multiplier with different types of reflex klystron, have been measured. Three types of multiplier are compared; these are (a) a commercial Si crystal mounted upright on the exterior face of the harmonic waveguide; (b) a crystal with windows cut in the face so that the tungsten whisker is exposed; (c) a whisker and Si block mounted in the waveguide. Arrangement (b) is the most satisfactory for wavelengths below 4 mm. The performance of arrangement (c) as a detector is discussed. A crystal harmonic generator used with a narrow-band amplifier will furnish enough power at 3-mm wavelength to give a dynamic range of 50–60 db.

621.373.424:621.396.621.54 2606

Variable-Frequency Crystal-Controlled Receivers and Generators—Wadley. (*See* 2759.)

621.373.431+621.373.444.1 2607

A Modified Miller Timebase Circuit—R. D. Ryan. (*Jour. Sci. Instr.*, vol. 31, pp. 73–75; March, 1954.) High linearity and rapid flyback are achieved by using a blocking oscillator to charge the feedback capacitor in the Miller-type circuit. Compensation is provided for variations in the blocking-oscillator characteristics. A unit giving sweep durations of 5 μs–15 ms with either triggred or continuous operation is described.

621.375.2.024:621.387 2608

Transient Response of Glow Discharges with Applications—Mackay and Morris. (*See* 2828.)

621.375.223 2609

L.F. Compensation for Video Amplifiers: Part 1—Amplifiers without Feedback—J. E. Flood. (*Wireless Eng.*, vol. 31, pp. 175–186; July, 1954.) Response to unit-step input is considered. The departure from the ideal flat-top response caused by low-frequency distortion can be reduced by making the differential coefficients of the output function equal to zero at $t=0$. The order of compensation is designated by the number of derivatives equal to zero. General theory is given and is applied to some typical resistance-coupled amplifiers. A single *rc*-coupled stage can have second-order compensation, and multistage amplifiers can have higher orders of compensation. Experimental results confirm the analysis.

621.375.23:621.317.44 2610

An Electronic Voltage Integrator—R. Madey and G. Farly. (*Rev. Sci. Instr.*, vol. 25, pp. 275–279; March, 1954.) A feedback-amplifier integrating circuit is described. The amplifier comprises a cascode-connected direct-coupled circuit with filament drift compensation and internal positive-feedback adjustment. The integrator is used with a search coil for magnetic-field measurements in the range 500–15000 gauss; accuracy is within 1 per cent.

621.375.23.029.4 2611

A High-Efficiency High-Quality Audio-Frequency Power Amplifier—A. B. Bereskin. (*Trans. I.R.E., I.R.E. Convention Record*, Part 6, pp. 18–24; 1954. pp. 49–60; March/April, 1954.) The circuit uses a Type-12AX7 tube as phase-inverter-amplifier-driver stage direct-coupled to two output beam tetrodes operating in class *B* push pull, with 24 db of feedback. The output transformer has a bifilar-wound primary, and the feedback winding is electrostatically shielded from the secondary but very closely coupled to it. Full-power output is 60w at frequencies <3000 cps. Design of the circuits and performance details are described.

621.375.23.029.4 2612

The Cascode as a Low-Noise Audio Amplifier—R. L. Price. (*Trans. I.R.E.*, vol. AU-2, pp. 60–64; 1954.)

621.375.3 2613

The Influence of Magnetic Amplifier Circuitry upon the Operating Hysteresis Loops—H. W. Lord. (*Trans. AIEE*, Part I, *Communication and Electronics*, vol. 72, pp. 721–728; 1953.) An analysis limited to the steady-state characteristics of single-phase self-saturated magnetic amplifiers. Only resistive loads are considered.

621.375.3 2614

High-Speed Magnetic Amplifiers—A. E. Maine. (*Elec. Eng.*, vol. 26, pp. 180–185; May, 1954.) Units based on Ramey's half-wave amplifier are discussed, and the development of a new series of amplifiers is described. The basic type is a reversing-phase-input/reversing-dc-output amplifier. Circuit modifications are explained for obtaining (a) reversing dc output for reversing dc input, and (b) reversing-phase ac or reversing polarity dc output. Other possible modifications are indicated.

621.375.3:621.318.1 2615

Theory of Magnetic Amplifiers with Square-Loop Core Materials—H. F. Storm. (*Trans. AIEE*, Part I, *Communication and Electronics*, vol. 72, pp. 629–637; 1953. Discussion, pp. 637–640.)

621.375.4:621.314.7 2616

Temperature-Stabilized Transistor Amplifiers—H. J. Tate. (*Electronics*, vol. 27, pp. 144–147; June, 1954.) Design equations are derived and monograms presented for determining the temperature variation of the operating point of a junction transistor amplifier. Application of the information for temperature stabilization is indicated. Numerical examples are given.

621.375.4:621.314.7 2617

Transistors in Amplifier Output Stages—M. J. O. Strutt. (*Scientia elect.*, *Zürich*, vol. 1, pp. 2–17; October, 1953.) The optimum effective load impedance and the efficiency are calculated using idealized characteristic curves. The results for point-contact and junction-type transistors in class A and in class B push-pull amplifiers are compared with the corresponding formulas for triodes and pentodes. Distortion, the efficiency limit, and circuit design are also briefly considered.

GENERAL PHYSICS

530.112 2618

Proposal for a New Aether-Drift Experiment—L. Essen. [*Nature* (*London*), vol. 173, p. 734; April 17, 1954.] Comment on Furth's proposal (1726 of June) and suggestion for an alternative experimental procedure.

535.22 2619

Precision Determination of the Velocity of Light Derived from a Band Spectrum Method: Part 2—D. H. Rank, J. N. Shearer, and T. A. Wiggins. (*Phys. Rev.*, vol. 94, pp. 575–578; May 1, 1954.) From measurements of the rotational constant B_{000} for HCN by the method of exact orders (*Jour. Opt. Soc. Amer.*, vol. 43, p. 952; 1953.) and by a microwave method, the value of *c* is 299789.8 ±3.0 km.

536.49:621.319.4 2620

Further Experiments on the Thermodielectric Effect (Costa Ribeiro Effect)—A. Dias Tavares. (*Ann. Acad. Bras. Sci.*, vol. 25, pp. 53–59; March 31, 1953. In English.) The experiments described establish that on solidification of liquid naphthalene, negative charges are expelled from the solid into the liquid phase, the resultant positive charge being distributed within the solidified dielectric. See also 648 of 1952 (Riberio).

536.49:621.319.4 2621

Further Quantitative Experiments on the Costa Ribeiro Effect—A. Dias Tavarse. (*Ann. Acad. Bras. Sci.*, vol. 25, pp. 353–373; December 31, 1953. In English.) In an investigation of the thermodielectric effect (2620 above and back reference), measurements were made of the charge within a solidified naphthalene dielectric by means of a Faraday cage and electrometer. The relation between the charge and the mass is constant. For amall crystals this specific residual charge ρ is twice the induced charge per unit mass. Impurities alter the value of ρ considerably.

537.221 2622

Contact Charging between a Borosilicate Glass and Nickel—J. W. Peterson. (*Jour. Appl. Phys.*, vol. 25, pp. 501–504; April, 1954.) Contact charging of glass spheres rolling on clean Ni was studied under conditions of controlled cleanliness, humidity, and gas pressure. The effect of a transverse electric field is important only for high surface conductivity. Surface conduction limits the maximum charge, which also varies with pressure due to gaseous discharge between the sphere and the metal.

537.311.33 2623

A Chemical Approach to the Treatment of Electronic Spin in Semiconductors—J. H. Crawford, Jr. and D. K. Holmes. (*Proc. Phys. Soc.*, vol. 67, pp. 294–295; March 1, 1954.) A method due to Fowler (*Proc. Roy. Soc. A*, vol. 140, p. 505; 1933.), in which the electrons and the neutral and ionized donors are treated as chemical entities, affords a simple approach in the nondegenerate case. The results obtained agree with those of Landsberg (78 of 1953) when the occurrence of neutral donors containing unpaired and paired electrons is taken into account.

537.311.62 2624

Theory of the Anomalous Skin-Effect in a Magnetic Field—M. Ya. Azbel' and M. I. Kaganov. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 95, pp. 41–44; March 1, 1954. In Russian.] Extension of work by Reuter and Sondheimer (1354 of 1949) to take account of the effect of a steady magnetic field.

537.311.62:535.137 2625

The Effect of Relaxation on Microwave Measurements of the Anomalous Skin Effect—R. G. Chambers. (*Physica*, vol. 19, pp. 365–370; April, 1953.) The ratio of the conductivity to the mean free path of an electron in Bi at microwave frequencies estimated on the basis of Dingle's theory (2626 below) is $1.53\times10^{7}\Omega^{-1}$ cm^{-2} compared with $2.74\times10^{7}\Omega^{-1}$ cm^{-2} given earlier by Pippard and Chambers (996 of 1953), neglecting relaxation effects. This discrepancy and the discrepancy for normal metals are discussed.

537.311.62:535.137 2626

The Anomalous Skin Effect and the Reflectivity of Metals—R. B. Dingle. (*Physica*, vol. 19, pp. 312–364, 729–736, and 1187–1199; April, August and December, 1953.) Expressions derived by Reuter and Sondheimer (1354 of 1949) are considerably simplified. Numerical results obtained include calculated values of surface resistance and reactance in the microwave region for both specular and diffuse electron reflection.

537.52 2627

Some Aspects of Breakdown Streamers—L. B. Loeb. (*Phys. Rev.*, vol. 94, pp. 227–232; April 15, 1954.)

537.523 2628

The Electric Strength of Air in Nonuniform Fields at Radio Frequencies—J. B. Whitehead, D. L. Birx, and C. F. Miller. (*Trans. AIEE*, Part I, *Communication and Electronics*, pp. 520–524; 1953.) Measurements were made of the breakdown potential of atmospheric air between a cylinder and an axial wire, over the frequency range 2–16 mc. The breakdown voltage is about 30 per cent below the 60-cps value at 2 mc, and decreases at the average rate of 120 per mc as the frequency is raised above this point. The results are explained by a modified Townsend theory. The discharge occurs during the part of the cycle when the wire is positive with respect to the cylinder.

537.525 2629

The Secondary-Electron Resonance Mechanism of Low-Pressure High-Frequency Gas Breakdown—A. J. Hatch and H. B. Williams. (*Jour. Appl. Phys.*, vol. 25, pp. 417–423; April, 1954.) Breakdown field strengths were measured in air and H_2 at frequencies from 25 to 90 mc between internal electrodes separated by 1–4 cm. By suddenly applying a high voltage, and reducing it slowly, an upper breakdown curve was observed. This curve, the lower breakdown curve and cut-off frequency data together define a breakdown region. The secondary-electron resonance theories developed by Danielsson, and Gill and von Engel (1906 of 1948) are extended.

537.525.3:621.387.424 2630

Dynamics of Corona Discharge between Cylindrical Electrodes—L. Colli, U. Facchini, E. Gatti, and A. Persano. (*Jour. Appl. Phys.*, vol. 25, pp. 429–435; April, 1954.) The screening action of the space charge is responsible for the constancy and stability of the average discharge current. Fluctuations are explained as a response of the system having a definite resonance frequency to the statistical fluctuations of the photoelectric current. Theoretical predictions are confirmed by experiment.

537.533 2631

The Work Function of Irregular Metal Surfaces T. J. Lewis. (*Proc. Phys. Soc.*, vol. 67, pp. 187–200; March 1, 1954.) Calculation of the image potential for spherical and prolate spheroidal bosses on an otherwise smooth surface shows that this potential might be reduced to 50 per cent of its value for an ideal plane surface. Since the image potential is likely to form a large portion of the work function for most metals, the work function will be significantly reduced when the surface is rough.

537.533.8 2632

Effect of the Energy Distribution of Electrons on the Average Secondary-Emission Yield of an Insulator—J. Salmon. (*Jour. Phys. Rad.*, vol. 15, p. 190; March, 1954.) An expression for the average yield in terms of electron temperature is calculated for pyrex, the electron energy distribution being Maxwellian.

537.56 2633

A Model for Collission Processes in Gases: Part 1—Small Amplitude Processes in Charged and Neutral One-Component Systems—P. L. Bhatnagar, E. P. Gross, and M. Krook. (*Phys. Rev.*, vol. 94, pp. 511–525; May 1, 1954.)

537.568 2634

Electron-Ion Recombination at Low Pressures—S. Borowitz. (*Trans. AIEE*, Part I, *Communication and Electronics*, vol. 72, pp. 430–435; 1953.) Experiments throwing light on the nature of the fundamental processes involved are discussed; the molecular ion is thought to play a decisive part. The evidence is consistent with the hypothesis of dissociative recombination.

538.114 2635

The Tensor Formulation of Ferromagnetic Resonance—J. A. Young, Jr. and E. A. Uehling. (*Phys. Rev.*, vol. 94, pp. 544–554; May 1, 1954.)

538.12:538.653.12 2636

Twisted Magnetic Fields in Conducting Fluids—J. W. Dungey and R. E. Loughhead. (*Aust. Jour. Phys.*, vol. 7, pp. 5–13; March, 1954.) The formation of loops in the lines of force of a twisted magnetic field confined within a cylinder of radius R is discussed by the method of normal modes. The condition obtained for loop formation is that the pitch of the twisted field be less than ΠR. The velocity of Alfvén waves in this model is also discussed.

538.248 2637

Diffusion After-Effect in Weak Alternating Fields. Validity of Rayleigh's Law—A. Marais. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 238, pp. 1782–1874; May 3, 1954.]

538.561:537.56 2638

On the Excitation of Oscillations in a Thermal Plasma—M. Sumi. [*Jour. Phys. Soc.* (*Japan*) vol. 9, pp. 88–92; January/February, 1954.] Plasma oscillations are considered for an arbitrary distribution of electron beams. The fundamental equations are derived assuming a weak signal and the absence of static fields. The dispersion relation derived is applied to the case of excitation of thermal plasma by an injected electron beam, and the rate of wave growth is determined. The numerical results are compared with results obtained by Merrill and Webb (*Phys. Rev.*, vol. 55, pp. 1191–1198, June 15, 1939).

538.566:535.42 2639

The Vector Wave Function Solution of the Diffraction of Electromagnetic Waves by Circular Disks and Apertures: Part 2—The Diffraction Problems—C. Flammer. (*Jour. Appl. Phys.*, vol. 25, p. 543; April, 1954.) Correction to paper abstracted in 708 of March.

539.145 2640

Determination of h/e by a New Method—F. G. Dunnington, C. L. Hemenway, and J. D Rough. (*Phys. Rev.*, vol. 94, pp. 592–598; May 1, 1954.) The method involves the determination of the excitation potential of *He* from the ground state to the lowest permitted singlet level. The value of h/e obtained is $(1.3790 \pm 0.0002) \times 10^{-17}$ erg. seconds/esu.

621.3.032.44 2641

Temperature Distribution in an Electrically Heated Filament—K. S. Krishman and S. C. Jain. [*Nature* (*London*), vol. 173, pp. 820–821; May 1, 1954.] A practical general solution is given for a filament of finite length heated in vacuo. See also 2375 of August.

537.533.8 2642

Physics and Applications of Secondary Electron Emission. [Book Review]—H. Bruining. Publishers: Pergamon Press, London, Eng., McGraw-Hill Book Company, Inc., New York, N. Y., 178 pp., 25s. or $5. (*Jour. Atmos. Terr. Phys.*, vol. 5, pp. 114–115; May, 1954.) The book "covers practically everything of importance, and should serve as an admirable guide to workers in all branches of electronics."

GEOPHYSICAL AND EXTRATERRESTRIAL PHENOMENA

523.7:550.38 2643

Relation between the Apperance of the Yellow Coronal Line and Geomagnetic Activity—J. F. Denisse and P. Simon. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 238, pp. 1775–1778; May 3, 1954.] From observations of the yellow coronal line it is possible to identify those solar-activity centers whose passage across the central meridian is accompanied by an increase of geomagnetic activity.

523.72:537.562 2644

Space-Charge-Wave Amplification in a Shock Front and the Fine Structure of Solar Radio Noise— H. K. Sen. (*Aust. Jour. Phys.*, vol. 7, pp. 30–35; March, 1954.) The dispersion equation corresponding to the non-Maxwellian velocity distribution of the particles in a shock front is derived. The roots indicate frequency bandwidths of space-charge wave amplification decreasing with the shock strength. It is suggested, in agreement with Denisse and Rocard (1900 of 1952), that the storm bursts of narrow bandwidth originating in shock fronts constitute the elementary fine-structure components of solar radio noise bursts.

523.72+523.74]:621.396.822 2645

Thermal Radio Emission from the Sun and the Source of Coronal Heating—J. H. Piddington and R. D. Davies. (*Mon. Not. R. Astr. Soc.*, vol. 113, pp. 582–596; 1953.) A statistical analysis is made of data obtained from measurements of solar radiation at frequencies between 600 and 9400 mc. The total radiation is separated into a component associated with sunspots and a "basic" component showing no such connection. Results indicate that (a) sunspot radiation and the very hot regions responsible for it diminish less rapidly than the spot area and exist for a month or more after the spot has vanished; (b) the basic component at some frequencies is much lower than previously believed; (c) *S*-regions above the sunspots are the source of coronal energy and gases spreading from them can completely replace the corona in about four days.

523.72:621.396.822 2646

Solar Radio Emissions at 4-Metres Wavelength during 1947–50 inclusive and their Relations to Solar Activity—J. S. Hey. (*Mon. Not. R. Aster. Soc.*, vol. 113, p. 653; 1953.) Summary of RRDE Rep. No. 372, October, 1952. Bursts at $4.1_{-m}\,\lambda$ of duration $>\frac{1}{4}$ minute and of intensity over a certain minimum, are analyzed for correlation with solar flares. The median lag of bursts behind associated flares is about a minute, but the dispersion of the time differences is such that an appreciable proportion of the bursts precede the flares. Secondary controlling factors seem to exist both for the solar phenomena and the related ionospheric phenomena. Analysis of the distribution of bursts confirms that there is a greater chance of bursts in association with flares on the eastern half of the sun's disk than on the western half.

523.72:621.396.822 2647

Observation of the Partial Solar Eclipse (February 14, 1953) at the Wavelength of 10 Centimeters (3000 Mc/s)—K. Aoki. [*Rep. Ionosphere Res.* (*Japan*) vol. 7, pp. 109–115; September, 1953.) The diameter of the sun observed at this wavelength is 5 per cent greater than that of the visual disk. Limb brightening is observed. Asymmetry in the eclipse curve is attributed to a local bright region round sunspots, where the values of electron density in chromosphere and corona are enhanced.

523.72:621.396.822 2648

The Distribution of Radiation across the Solar Disk at Metre Wavelengths—P. A. O'Brien. (*Mon. Not. R. As.r. Soc.*, vol. 113, pp. 597–612; 1953.) Measurements were made at wavelengths of 1.4 3.7, and 7.9 m, using variable-aperture interferometers and phase-switching receiver systems. The size of the emitting disk increases with increasing wavelength. Experiments with interferometers inclined at various angles to the solar axis of rotation showed the radiating shape of the sun to be elliptical; the radial distance at which the brightness temperature was reduced to half was about 25 per cent greater at the equator than in the polar direction.

523.72:621.396.822 2649

Distribution of Radio-Frequency Brightness across the Solar Disk and the Derivation of a Model Corona—P. A. O'Brien and C. J. Bell. [*Nature* (*London*), vol. 173, p. 219; January 30, 1954.] The derivation of a model consistent with observations previously reported (2648 above) is discussed.

523.72:621.396.822.029.62 2650

Solar Radio Asymmetry at 4-Metres Wavelength—J. S. Hey and V. A. Hughes. [*Nature* (*London*), vol. 173, p. 771; April 24, 1954.] The greater frequency of observation of solar bursts at 4.1 m in association with flares on the eastern half of the sun's disk [1368 of 1949 (Hey et al)] may be due to absorption of bursts from the western half in ionized streams. This would account for the lack of evidence of asymmetry at 1.5 m, for which wavelength absorption is much less. An analysis of this asymmetry with reference to geomagnetic activity is being made.

523.852:621.396.822 2651

Intensities of Discrete Radio Sources in Cygnus and Cassiopeia at 22.6 Mc/s—J. S. Hey and V. A. Hughes. [*Nature* (*London*), vol. 173, pp. 819–820; May 1, 1954.] Measured intensities are 9.4×10^{-22} and 4.6×10^{-22} W/m^2.(cps) for the sources in Cassiopeia and Cygnus respectively. Comparison with intensities measured previously at higher frequencies indicates that in the spectrum of the Cassiopeia source intensity is proportional to (frequency)$^{-1.2}$, the variation in the Cygnus source being similar except at lower frequencies where the rate of increase of intensity is reduced.

523.852.3:621.396.822 2652

Radio Emission from the Perseus Cluster—J. E. Baldwin and B. Elsmore. [*Nature* (*London*), vol. 173, p. 818; May 1, 1954.] Results of observations at $3.7_{-m}\,\lambda$, using radio-interferometer apertures of 14 λ and 157λ, provide confirmation for the identification of the radio source with *NGC* 1275.

523.854:621.396.822 2653

Galactic Radiation at Radio Frequencies: Part 7—Discrete Sources with Large Angular Widths—J. G. Bolton, K. C. Westfold, G. J. Stanley, and O. B. Slee. (*Aust. Jour. Phys.*, vol. 7, pp. 96–109; March, 1954.) Observations with a 72-foot reflector at a frequency of 150 mc, with the sea interferometer at 110 mc, and with the azimuth interferometer at 100 mc, revealed the existence of a number of sources of angular width >1 degree. The observations are described and discussed; the limitations of interference techniques are summarized. Part 6: 1756 of June (Bolton, Slee and Stanley).

523.854:621.396.822 2654

Galactic Radiation at Radio Frequencies: Part 8—Discrete Sources at 100 Mc/s between Declinations +50° and −50°—J. G. Bolton, G. J. Stanley, and O. B. Slee. (*Aust. Jour. Phys.*, vol. 7, pp. 110–129; March, 1954.) One hundred and four discrete sources have been found. Individiual sources are compared in position and flux density with those of previous surveys. The observed distribution of sources is discussed. Several identifications with visible sources are suggested. Part 7: 2653 above.

523.854:621.396.822 2655

A Comparison of the Intensities of Cosmic Noise observed at 18.3 Mc/s and at 100 Mc/s—C. A. Shain. (*Aust. Jour. Phys.*, vol. 7, pp. 150–164; March, 1954.) The principal conclusion of this comparison is that the background radiation cannot be made up of the radiation from sources of the type observed so far, although a small polar component may be due to the extragalactic sources. Absorption in interstellar gas has a considerable effect on the intensity variations of background radiation with direction observed at 18.3 mc.

523.854:621.396.822:550.510 535 2656

Observations of the General Background and Discrete Sources of 18.3-Mc/s Cosmic Noise—C. A. Shain and C. S. Higgins. (*Aust. Jour. Phys.*, vol. 7, pp. 130–149; March, 1954.) A survey of a broad strip of sky, centered on declination —32 degrees, has been made using an antenna with beam width to half power of 17 degrees. Previous results concerning the background distribution of brightness have been confirmed and 37 discrete sources have been detected. The distribution of these sources is shown. No correlation was found between the occurrence of scintillations and published ionospheric data, but the observations are consistent with an origin of scintillations in irregularities of dimensions about 4 km at a height of about 500 km.

523.854:621.396.822.029.62 2657

Galactic Radio Sources of Large Angular Diameter—R. H. Brown, H. P. Palmer, and A. R. Thompson. [*Nature* (*London*), vol. 173, pp. 945–946; May 15, 1954.] Preliminary results are given of measurements of the apparent angular width of sources previously observed [1408 of May (Brown and Hazard)].

523.854.22:621.396.822:621.396.677.3 2658

Wide-Band Radio Interferometer—V. V. Vitkevich. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 91, pp. 1301–1303; August 21, 1953. In Russian.] The use of a wideband amplifier, with the frequency characteristic given, in conjunction with an antenna system consisting of two or four antennas in line, or 16 antennas arranged symmetrically in a square, results in very high resolution. In the example quoted radio sources separated by an angle of 15′ can be resolved. The theory of the interferometer is given and the general form of the directional characteristic is shown.

550.384 2659

Geomagnetic Bay Disturbances and their Nonuniform Induced Components within the Earth—H. Wiese. (*Z. Met.*, vol. 8, pp. 77–79; February/March, 1954.) The bay disturbance is considered as the field of a traveling dipole in the auroral zone, the direction of travel depending on time of day. The induced component within the earth exhibits local variations depending on the conductivity; a "dynamic" anomaly is observed in an area in Central Europe, the vertical component being a maximum in the north, and a minimum in the south, when the ionosphere current is directed southward.

550.385/.386:525.233 2660

The Use of Earth-Potential Measurements for Magnetic- and Ionospheric-Storm Indication—A. H. DeVoogt. (*Jour. Atmos. Terr. Phys.*, vol. 5, pp. 108–110; May, 1954.) Using the underground telephone cable system, potentials are measured in directions as nearly parallel and perpendicular to the magnetic meridian as possible. Each of the cable circuits is over 100 km long and is provided with several earthing points at the ends. The recorders are housed in the control room of the PTT station Radio-Kootwijk.

551.5:621.396.11.029.6:061.3 2661

Radio-Meteorology: Conference in Texas—Saxton. (*See* 2751.)

551.510.3:535.325 2662

A Preliminary Survey of Tropospheric Refractive Index Measurements for U. S. Interior and Coastal Regions—C. M. Crain, J. R. Gerhardt, and C. E. Williams. (*Trans. I.R.E.*, vol. AP-2, pp. 15–22; January, 1954.) Results of some 700 recordings made between July and December, 1952, using airborne refractometers, at heights up to 25,000 feet, are summarized. The existence of large differences in refractive index near certain cloud boundaries is confirmed. See also 1759 of June (Crain et al).

551.510.535 2663

Atmospheric Space Charge—W. D. Parkinson: J. A. Chalmers. (*Jour. Atmos. Terr. Phys.*, vol. 5, pp. 106–108; May, 1954.) Comment on 131 of January and author's reply.

551.510.535 2664

The Evaluation of Ionospheric Observations—G. Grawert and H. Lassen. (*Z. angew. Phys.*, vol. 6, No. 3, pp. 136–143 March 1954.) A critical discussion of the various methods used in calculating the height distribution of the charge-carrier concentration and other parameters, from experimentally determined $h'f$ curves. The corrections and the estimation of errors in the calculated *F*-layer height in the presence of an *E* layer and the effect of neglecting the *E* layer in this calculation are discussed. An approximate method is indicated for taking the geomagnetic field into account.

551.510.535 2665

Semidiurnal Currents and Electron Drifts in the Ionosphere—J. A. Fejer. (*Jour. Atmos. Terr. Phys.*, vol. 5, p. 103; May, 1954.) Correction to paper abstracted in 1420 of May.

551.510.535 2666

A Subsidiary Layer in the E Region of the Ionosphere—R. Naismith. (*Jour. Atmos. Terr. Phys.*, vol. 5, pp. 73–82; May, 1954.) Echoes from sporadic-*E* can be distinguished from those from a layer at 90–100 km height. It is suggested that the ionization in this layer results from the impact of meteors on the atmosphere and that it may therefore be called the meteoric *E*-layer. The distinctive properties may be used to extend the use of the ionosphere for intermediate-distance radio communication.

551.510.535 2667

Origin of the Ionospheric E Layer—K. Rawer and É. Argence. (*Phys. Rev.*, vol. 94, pp. 253–256; April 15, 1954.) Critical discussion of theories of formation based on (a) ionization of O_2 by ultraviolet light, and (b) ionization by solar soft X-rays. The intensity of incoming radiation seems insufficient, in view of absorption in case (a), but from Elwert's computations (*Z. Naturf.*, vol. 7(a), pp. 202–204; 1952), sufficient in case (b). Provided dissociation of O_2 occurs in a high transition layer of considerable thickness, both processes may be involved, but process (b) seems the more important.

551.510.535 2668

Multiple Stratification of the F Layer at Ibadan—N. J. Skinner, R. A. Brown and R. W. Wright. (*Jour. Atmos. Terr. Phys.*, vol. 5, pp. 92–100; May, 1954.) Observations made during the period December 1951–January 1953 are reported. In the F_2 layer there is a maximum occurrence of ridges around 1,000 and 1,500 hours, the first maximum being the more pronounced. A seasonal variation is found, a maximum number of ridges being observed when the midday minimum of f_0F_2 occurs early in the day. A lunar semidiurnal variation is detectable during the morning hours. Formation of ridges is also observed in the F_1 layer, but there is little relation to those of the F_2 layer. Possible phases of vertical drift velocities that could explain the observations are discussed.

551.510.535 2669

On the F₂ Layer Distribution in Polar Region—T. Obayashi. [*Rep. Ionosphere Res.* (*Japan*), vol. 7, pp. 118–121; September, 1953.] Results of an analysis of monthly mean values of f_0F_2, $h'F_2$, f_0F_1 and f_0E for March 1951 are summarized. In a region along the auroral zone where the $h'F_2$ values are anomalously high, maxima occur at noon and midnight, and the region rotates round the earth with the sun. The increase of apparent height of the F_2 layer is due to wave retardation in the F_1 layer during daytime in high latitudes.

551.510.535 2670

A Theory of Distribution and Variation of the Ionospheric F₂ Layers—K. Maeda. [*Rep. Ionosphere Res.* (*Japan*), vol. 7, pp. 81–107; September, 1953.] "A brief description is first given on the results of our recent studies, which have led to the conclusions of inhibition of vertical ionospheric current, enhancement of conductivity, dynamo current, and vertical drift of charged particles near the geomagnetic equator and suppression of daytime electron density of F_2 layer caused by an upward drift of electrons. The treatment is extended to the case of ionospheric storm accompanied with geomagnetic storm. The comparison of the theoretical and observational results is made for the cases of undisturbed as well as disturbed states of F_2 layer. The agreement is generally good for undisturbed case, but not quite satisfactory for disturbed case. The mechanism of dynamo in the ionosphere is discussed and a hypothetical consideration concerning a wind system during a storm is given."

551.510.535:523.3:621.396.11 2671

Lunar Radio Echoes and the Faraday Effect in the Ionosphere—Murray and Hargreaves. (*See* 2748.)

551.510.535:523.75 2672

The Enhancement of Ionospheric Ionization during Solar Flares—A. P. Mitra and R. E. Jones. (*Jour. Atmos. Terr. Phys.*, vol. 5, pp. 104–106; May, 1954.) Study of the recombination coefficient in the lower ionosphere, based on the decay of sudden phase anomalies at 16 kc and on sudden enhancements of atmospherics, gives values much smaller than those reported by Bates and Seaton (1407 of 1950). Measurements of the delay between the time of maximum flare and the times of maxima of sudden phase anomalies and sudden enhancements of atmospherics also suggest smaller values. These results indicate that for any height the value of the recombination coefficient remains unaltered during a sudden ionospheric disturbance and that the sudden increase in electron concentration is due to an increase in the rate of electron production in the layer.

551.510.535:523.78 2673

Corpuscular Eclipse in the F₂ Layer and its Association with Solar Flares and M-Regions—P. K. S. Gupta and S. N. Nitra. [*Nature* (*London*), vol. 173, pp. 814–816; May 1, 1954.] A survey of present knowledge on solar sources of corpuscular streams and on corpuscular velocities. The drop in the F_2-layer ion density occurring approximately five hours before the maximum of the optical solar eclipse is briefly discussed and it is suggested that in future F_2-layer observations should commence at least seven hours before the moment of totality.

551.510.535:621.396.11.029.53 2674

High Multiple Radio Reflections from the F_2 Layer of the Ionosphere at Brisbane—Baird. (*See* 2750.)

551.593 2675

The Effect of Rayleigh Scattering and Ground Reflection upon the Determination of the Height of the Night Airglow—E. V. Ashburn. (*Jour. Atmos. Terr. Phys.*, vol. 5, pp. 83–91; May, 1954.)

LOCATION AND AIDS TO NAVIGATION

621.396.9:621.396.621 2676

Aircraft Receiver for VOR-ILS and Communications—G. W. Gray. (*Electronics*, vol. 27, pp. 180–184; June, 1954.) Description of complete navigation and communication equipment. Cro presentation of omnirange signal and cr indicator for instrument landing system right-left signals are provided. High sensitivity is achieved in the omnirange receiver by using the extremely low bandwidth of about 6 cps in order to reduce noise. Special arrangements are made to avoid errors due to possible phase shift resulting from use of this narrow band.

621.396.93+621.398:621.396.665 2677

Sequentially Gated Automatic Gain Control—M. Eliason. (*Electronics*, vol. 27, pp. 186–188; June, 1954.) The arrangement described, which is applicable to navigation and telemetry systems, supplies a control voltage which holds receiver output constant within ±1 db when pulse signals are received from two or more transmitters in turn on the same frequency but with widely different strengths. Separate agc circuits corresponding to each transmitter are provided and are gated sequentially in synchronism with the transmitters.

621.396.933:621.396.621 2678

Airborne Loran Receiver—the AN/APN-70—R. R. Freas. (*Trans. I.R.E.*, vol. ANE-1, pp. 17–25; March, 1954.) The AN/APN-70 equipment was designed primarily for improved operating convenience and reliability as compared with older types. Improvements introduced include (a) time-difference information presented directly on counter dials, (b) possibility of operation on the new lf channels, giving instantaneous indication of position, (c) afc of the timing oscillator, (d) independent gain control for signals from each ground station, (e) high-sensitivity antenna coupler, and (f) all circuits and components selected for their reliability.

621.396.96:621.316.726 2679

Automatic Tuning for Primary Radar—S. Ratcliffe. (*Wireless Eng.*, vol. 31, pp. 122–131, 153–165, and 187–192; May/July, 1954.) Research into the basic techniques of automatic frequency correction is reported and design principles are discussed with particular attention to the requirements of airborne microwave radar equipment. For convenience, the frequency control is applied to the receiver rather than the transmitter, maintenance of the desired IF being more important than the absolute frequency. The frequency variations to be controlled may be either long-term, such as those due to changes in ambient temperature or pressure, or short-term, such as variations in magnetron freqency due to changes in the matching during rotation of the scanner. Receiver bandwidth and sensitivity are considered in relation to the conditions under which it may be economical to use an afc system with a response rapid enough to deal with scanner pulling. Circuits with the smallest possible number of pre-set adjustments should be used; adoption of a wide-band afc system is advantageous from this point of view.

621.396.93:621.396.676 2680

U.H.F. Omnidirectional Antenna Systems for Large Aircraft—Sichak and Nail. (*See* 2568.)

621.396.962.3 2681

Improved Demodulator for Radar Ranging—C. E. Goodell. (*Electronics*, vol. 27, pp. 170–171; June, 1954.) Gaps due to loss of echo pulses as a result of scattering, absorption, etc. are filled in by pulses generated in the demodulator. Range errors are thus reduced.

621.396.965 2682

Antenna Scan Considerations—D. Levine. (*Trans. I.R.E.*, vol. ANE-1, pp. 26–40; March, 1954.) Three types of scan are considered,(a) the spiral scan, (b) the simple conical scan, (c) sector scanning, the antenna having simple harmonic motion. In general a harmonic type of sector scan is to be preferred to a uniformly rotating system of n antennas, provided the total angle of scan is <229.2 degrees/n. A figure of merit for a sector scan having nonharmonic motion is also derived.

MATERIALS AND SUBSIDIARY TECHNIQUES

531.788.7 2683

Pressure Measurement with the Pirani Gauge—K. D. Mielenz. (*Z. angew. Phys.*, vol. 6, pp. 101–104; March, 1954.) The design of a Pirani gauge is discussed on the basis of expressions for the thermal losses of the heater wire and for the sensitivity of the gauge when used in a Wheatstone bridge. For maximum sensitivity the requirements include a galvanometer with high voltage sensitivity, low-resistance bridge arms, a low external temperature obtained by immersing the gauge in liquid air, and low heater-wire temperature obtained by using high-resistivity wire. These requirements were confirmed experimentally by measurements at pressures down to $1.7+10^{-5}$ Torr.

531.788.7 2684

A Simple Thermionic Vacuum Gauge—G. K. T. Conn and H. N. Daglish. (*Jour. Sci. Instr.*, vol. 31, pp. 95–96; March, 1954.) The gauge described is similar in essentials to that of Bayard and Alpert (2785 of 1950), but is much easier to construct.

533.583 2685

Influence of Electronic Impact on the Rate of Sorption of Gases on to Getter Materials—S. Wagener. [*Nature* (*London*), vol. 173, pp. 684–685; April 10, 1954.] Experiments made with Ba and Mg getters, using a Knudsen gauge and a special diode to produce electron impact, indicate that the introduction of the latter does not materially affect the gettering rate.

535.215:538.632:546.311.12 2686

Electronic Hall Effect in the Alkali Halides—A. G. Redfield. (*Phys. Rev.*, vol. 94, pp. 537–540; May 1, 1954.) The Hall mobility of electrons in NaCl, KCl, KBr and KI was measured using the technique described in 2687 below.

535.215:538.632:549.211 2687

Electronic Hall Effect in Diamond—A. G. Redfield. (*Phys. Rev.*, vol. 94, pp. 526–537; May 1, 1954.) The crystal is placed between two glass plates coated with a very thin semiconducting film. An electrometer is connected to the upper plate and the potential of the lower plate is varied. To make a measurement the magnetic field is applied in opposite directions alternately, the crystal being illuminated and the potentiometer adjusted to give zero deflection on the electrometer for each direction. The Hall angle can be obtianed approximately from the measured voltage change and the electrode geometry. Sources of error and corrections are considered in detail. Electronic mobility varies as $T^{-3/2}$ and is about 1,800 cm/s per V/cm at 300 degrees K, the corresponding figures for holes being $T^{-3/2}$ and 1,200 cm/s per V/cm.

535.215:546.289 2688

Photoconductivity in Gold-Germanium Alloys—R. Neuman. (*Phys. Rev.*, vol. 94, pp. 278–285; April 15, 1954.) Photoconductivity in n- and p-type gold-doped Ge was investigated at 77 degrees and 22 degrees K. The low-energy thresholds for impurity photoconduction are compatible with the activation energies measured by electrical methods. In some n-type specimens quenching effects were observed with prominent quench bands at 0.8 and 0.66 eV. Characteristics of the quenching were measured as functions of light intensity, applied voltage and temperature. No complete explanation is offered.

535.215:546.431-31:539.234 2689

External Photoemission of Thin Films Deposited by Evaporation from Barium Oxide—S. Narita. [*Jour. Phys. Soc.* (*Japan*), vol. 9, pp. 22–27; January/February, 1954.] Experimental investigation of films deposited on Mo and on Ta surfaces is reported. The variation of the energy distribution of the photoelectric emission and the thermionic work function with layer thickness is shown graphically and the change from metallic-type photoelectric emission, expressed by Fowler's equation, to the semiconductor type is noted. The dependence of photoelectric emission on the plane of polarization of the incident light is also shown graphically.

535.37 2690

Comparative Measurements of Electron-Excited Luminescent Screens for Front Viewing and for Back Viewing—B. Deubner and F. Hieber. (*Z. angew. Phys.*, vol. 6, pp. 112–115; March, 1954.) Light-yield measurements were made on Cu-activated ZnS screens excited by a 40-kV electron beam incident at 45 degrees. Screens of various thicknesses were used, with and without an Al base. Maximum yields in transmission-type screens, equal to those of front-viewing types, were obtained using an Al layer separated by collodion from the luminescent material. The resolution at various grain sizes was also measured. The results are shown graphically.

535.37:546.817.221 2691

Photoluminescence of PbS in the Infrared Region of the Spectrum—L. N. Galkin and N. V. Korolev. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 92, pp. 529–530; September 21, 1953. In Russian.] Brief account of method and results of an investigation of photoluminescence in the region 2–3.5 μ.

537.226 2692

Photodielectric Effect [dielectric loss change]—J. Roux. (*Jour. Phys. Rad.*, vol. 15, pp. 176–188; March, 1954.) A critical survey of work done on this subject. Most crystalline phosphors exhibit the effect, the magnitude of which depends largely on the phosphor type and preparation, and is greatly influenced by the nature and intensity of the excitation radiation, and by the intensity and frequency of the electric field applied to measure the capacitance of the material. A decay in the effect closely following the decay in phosphorescence is observed. Some relation between phosphorescence and dielectric loss change exists, but its exact nature has not been established.

537.226:538.56.029.4/.6 2693

The Radio Spectrum and Structure of Solids: Part 1—Debye Absorption in Solids, at Frequencies within the Radio Spectrum, and Lattice Defects—M. Freymann and R. Freymann. (*Jour. Phys. Rad.*, vol. 14, pp. 203–211; March, 1953.) In the study of solids by examination of their dielectric properties, two types of dipole must be distinguished, (a) dipoles associated with orientation of molecules or groups of atoms, and (b) dipoles associated with lattice defects. Experimental methods used to measure absorption at frequencies between 300 kc and 300 kmc are surveyed.

537.226:538.56.029.4/.6 2694

The Radio Spectrum and Structure of Solids: Part 2—Debye Absorption in Water in the Free and in the "Bound" State—M. Freymann and R. Freymann. (*Jour. Phys. Rad.*, vol. 15, pp. 165–175; March, 1954.) Critical examination of experimental data with a view to extending the theory developed in Part 1 (2693 above) to ionic crystals and semiconductors.

537.311.3:539.23 2695

Tentative Explanation of the Conduction Mechanism of Thin Granular Metal Layers—N. Nifontoff. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 238, pp. 1870–1872; May 10, 1954.] Experimental results on the variation of the resistance of granular metal layers can be explained by assuming the existence of a more or less continuous impurity-rich semiconductor layer between the grains. An electron passing from one grain to another then crosses two metal-semiconductor boundaries, the resistance being localized almost entirely at the reverse boundary.

537.311.3:539.23 2696

High-Frequency Resistance of Thin Films—R. Broudy and H. Levinstein. (*Phys. Rev.*, vol. 94, pp. 285–289; April 15, 1954.) Measurements on evaporated Ge, Rh and Au films show that for uniform films the decrease of resistance with frequency is due to simple self-capacitance. When the ratio of the ac and dc resistances is plotted against the product of frequency and dc resistance, data for uniform films of each material fall on a single curve. The presence of nonuniformities in the film produces an additional decrease of resistance with frequency corresponding to the degree of variation in resistance per unit length.

537.311.3:539.23:621.383.4 2697

High-Frequency Resistance of Photoconducting Films—R. Broudy and H. Levinstein. (*Phys. Rev.*, vol. 94, pp. 290–292; April 15, 1954.) PbTe photocells were prepared by evaporation under controlled conditions, and the resistance/frequency characteristic measured for various types of illumination. The degree of irregularity in response correlated with the degree of nonuniformity of the film, the nonuniformity detected decreasing on exposure to light.

537.311.33 2698

Semiconducting Compounds and the Scale of Electronegativities—C. H. L. Goodman. (*Proc. Phys. Soc.*, vol. 67, pp. 258–259; March 1, 1954.) In considering the relation between the ionic component of bonding and energy gap, suggested by Welker (1798 of June) a useful guide to the ionicity of a binary compound can be obtained from the electronegativity difference of its element. Correlations between energy gap and electronegativity difference exists for certain compounds between groups III and V and between groups II and VI of the periodic table. Mobility values for HgSe, InSb and HgTe support this hypothesis. HgSe may be of particular interest for semiconducting devices such as crystal triodes.

537.311.33:546.23 2699

Experimental and Theoretical Investigation of the Isothermal and the Adiabatic Hall Effect in Se—R. Diestel. (*Z. Naturf.*, vol. 8(a), pp. 453–457; August, 1953.) The isothermal and adiabatic Hall coefficients are calculated for a *p*-type semiconductor. The difference between the two values decreases rapidly with increasing thermal conductivity of the lattice, but even for substances, such as Se, with low thermal conductivity the difference amounts to only about 1 per cent. This result is confirmed by measurements on Se.

537.311.33:546.23 2700

Hole Conduction in Crystalline Selenium—W. Schottky. (*Z. Naturf.*, vol. 8(a), pp. 457–459; August, 1953.) Results obtained by Diestel (2699 above) indicate that the value of the thermoelectric force for Se found from direct measurement is more than twice the value found from Hall-effect measurements. It is tentatively suggested that the discrepancy can be explained by assuming an impurity-band conduction mechanism rather than conduction by free holes.

537.311.33:[546.24+546.3-1-24-23 2701

Electrical Properties of Pure Tellurium and Tellurium-Selenium Alloys—A. Nussbaum. (*Phys. Rev.*, vol. 94, pp. 337–342; April 15, 1954.) Measurements of resistivity and Hall coefficient for pure single-crystal Te and for six different single-crystal TeSe alloys with a maximum Se content of about 15 per cent by weight were made over the temperature range 90 degrees and 550 degrees K. An increase in the forbidden-band width and a lowering of the upper anomalous Hall-coefficient reversal temperature with increased Se content was observed.

537.311.33:546.24.03 2702

Electrical Properties of Tellurium Crystals at Very Low Temperatures—T. Fukuroi, S. Tanuma, and Y. Muto. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. A*, vol. 6, pp. 18–29; February, 1954.) The electrical conductivity, the Hall effect and the magnetoresistance effect were investigated in the temperature range from 1.7 degrees to room temperature, using both pure single crystals of Te and Te crystals including 0.01 per cent Sn. The results are tabulated and shown graphically.

537.311.33:546.28 2703

Hall Mobility of Electrons and Holes in Silicon—P. P. Debye and T. Kohane. (*Phys. Rev.*, vol. 94, pp. 724–725; May 1, 1954.) Mobility /resistivity curves are shown, obtained at 295 degrees K on single-crystal specimens of *n* and *p* type ranging in resistivity from 0.01 to 94 Ω.cm and from 0.025 to 110Ω.cm respectively. For the highest resistivity values the mobility for electrons is about 1,900 cm/s per V/cm and for holes 425 cm/s per V/cm.

537.311.33:546.289 2704

Resistivity Striations in Germanium Crystals—P. R. Camp. (*Jour. Appl. Phys.*, vol. 25, pp. 459–463; April, 1954.) Variations in conductivity may be made visible by electroplating, with CuSO$_4$ as electrolyte, since more copper will be deposited on regions having low resistivity. Pulsed voltage is applied. Maximum visual contrast is obtained by suitably adjusting the resistivity of the electrolyte, the magnitude and duration of the pulses and the pulse repetition frequency.

537.311.33:546.289:621.396.822 2705

Noise in Germanium Filaments at Very Low Frequencies—B. V. Rollins and I. M. Templeton. (*Proc. Phys. Soc.*, vol. 67, p. 271; March 1, 1954.) Measurements using the technique previously proposed (2039 of 1953) are reported.

537.311.33:546.482.21.03:539.234 2706

The Optical and Electrical Properties of Cadmium Sulphide Films—G. Kuwabara. [*Jour. Phys. Soc.* (*Japan*), vol. 9, pp. 97–102; January/February, 1954.] An experimental investigation of evaporated films shows that the optical properties are reproducible and similar to those of CaS single crystals, while the electrical properties depend on the conditions of evaporation and after-treatment. A close correlation between dark conductivity and photosensitivity was found. The results are shown graphically and are discussed.

537.311.33:546.682.86 2707

Electrical Properties of InSb—H. Weiss. (*Z. Naturf.*, vol. 8(a), pp. 463–469; August, 1953.) Continuation of work described by Welker (1798 of June). Measurements of conductivity, Hall effect and magnetoresistance are reported. Both *n*- and *p*-type samples were observed, depending on the impurity concentration. At room temperature, the highest value of electron mobility found was 41,000 cm/s per V/cm, and the greatest variation of resistance in a magnetic field of 10,000 gauss amounted to 150 per cent.

537.311.33:548.5 2708

Crystal Growth by the Tip Fusion Method—P. H. Keck, S. B. Levin, J. Broder, and R. Lieberman. (*Rev. Sci. Instr.*, vol. 25, pp. 298–299; March, 1954.) The main feature of the method described is the fusion of the tip of the boule into a sessile drop by means of a tungsten ring heated by hf induction. A continuous stream of powder is supplied from above, and the growing crystal is withdrawn slowly downward from the hot zone. Polycrystalline Si boules about 1 inch long and $\frac{1}{4}$ inch in diameter were grown at a rate of 1 inch per hour.

538.221:539.234 2709

The Magnetization of Very Thin Iron Films—W. Reincke. (*Z. angew. Phys.*, vol. 137, pp. 169–174; March 15, 1954.) Iron films evaporated on to glass plates were suspended by spun thread in a magnetic field and set in torsional oscillation. Susceptibility was calculated from the oscillation period, the field strength, the moment of inertia of the plates, and the volume of the film. Film thickness *d* was measured optically. Results are in agreement with calculations of Klein and Smith (1675 of 1951). Magnetization starts to decrease at a value $d=125$ Å, disappearing at $d=11$ Å. For $d<100$ Å saturation is reached at much lower field strengths than are required for thicker films.

538.221:621.318.134 2710

Magnetic-Dispersion Spectrum of a Ni-Zn Ferrite—L. Lucas. (*Z. angew. Phys.*, vol. 6, pp. 127–130; March, 1954.) The decrease of permeability with increasing frequency is explained on the basis of superposition of sharp spin-resonances. From the distribution function of characteristic frequencies, approximate formulas are derived for the observed relation between the static susceptibility and the lower limiting characteristic frequency, the dispersion of the loss factor at low frequencies and the decrease of permeability at high frequencies. The width of the dispersion spectrum is determined by the internal demagnetization at the Weiss-domain boundaries.

538.221:621.318.134 2711

Magnetic Behaviour of some Ferrites—K. F. Niessen. (*Physica*, vol. 19, pp. 1127–1132; November, 1953.) Calculations based on a series of measurements on mixed crystals of nickel ferrite and nickel titanate, Fe$_{2-2a}$Ni$_{1+a}$Ti$_a$O$_4$ with $0<a<0.5$, show that the Ti appears to be distributed statistically amongst the tetrahedral and octahedral sites of the spinel when $a \ll 1$ but appears to show preference for the tetrahedral sites as *a* increases.

538.221:621.318.134.029.64/.65 2712

Resonance Absorption in Ferrites—T. Okamura. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. A*, vol. 6, pp. 89–108; February, 1951.) An experimental investigation over the frequency range 9.45 kmc–47 kmc at room temperature and at −195 degrees C. is reported. The real spectroscopic splitting factor, *g*, which is independent of frequency and temperature, was determined for ferrites of Ni, Mn, Ni-Zn and Mn-Zn by using the corrected Kittel resonance formula. A semi-empirical formula for the internal field is given. Results for true *g*-values and the internal fields are tabulated and the latter compared with calculated values.

538.221:621.318.134.029.64/.65 2713

The Effect of the Anisotropy and the Relaxation Phenomena on the Shift of Ferromagnetic Resonance in Polycrystalline Ferrites—M. Date. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. A*, vol. 6, pp. 109–113; February, 1954.)

548.55:546.78 2714

Thermionic and Surface Properties of Tungsten Crystals—G. F. Smith. (*Phys. Rev.*, vol. 94, pp. 295–308; April 15, 1954.) The relation between the production process for single crystals and types of surface structure was studied, and thermionic constants were determined for various crystallographic directions.

548.55:546.833 2715

Growth and Surface Properties of Tantalum

Crystals—M. H. Nichols. (*Phys. Rev.*, vol. 94, pp. 309–313; April 15, 1954.) A study of Ta on the lines of 2714 above.

621.315.612.4:546.431.824–31 2716

Anomalous Temperature Coefficient of Permittivitiy in Barium Titanate—K. W. Plessner and K. A. Cook. [*Nature* (*London*), vol. 173, pp. 682–683; April 10, 1954.] A negative temperature coefficient of permittivity was observed in a very pure ceramic $BaTiO_3$ sample subjected to alternations of temperature around the Curie point. The anomaly is related to the existence of the cubic and tetragonal phases. The observations support the view that the transition at about 120 degrees C. is a first-order transition.

MATHEMATICS

519.2 2717

Statistics of a Population with Creation and Recombination Dependent on Existing Numbers—D. A. Bell. (*Proc. Phys. Soc.*, vol. 67, pp. 227–231; March 1, 1954.) "Populations liable to loss by recombination are likely to have a rate of loss proportional to the square of the number present; and the distribution function is then related either to a Bessel function or to a Laguerre polynomial, according as the rate of creation is either constant or dependent on the complement of the number already present." Statistics of the number of particles present are then no longer the same as those of the density fluctuations in a perfect gas.

MEASUREMENTS AND TEST GEAR

621.3.014(083.74) 2718

Preliminary Development of a Magnettor Current Standard—E. P. Felch and J. L. Potter. (*Trans. AIEE*, Part I, *Communication and Electronics*, vol. 72, pp. 534–531; 1953.)

621.316.84(083.74) 2719

The Behaviour of Gold-Chromium Standard Resistors under Load—A. Schulze and D. Bender. (*Z. angew. Phys.*, vol. 6, pp. 132–136; March, 1954.) The effect of loads up to 1w on the resistance of 1-Ω, 10-Ω and 100-Ω resistors was measured; the results are tabulated.

621.317.335:537.228.1 2720

Methods for Measuring Piezoelectric, Elastic, and Dielectric Coefficients of Crystals and Ceramics—W. P. Mason and H. Jaffe. (PROC. I.R.E., vol. 42, pp. 921–930; June, 1954.) Various methods are reviewed, with a view to standardization. For investigating small specimens, the click-oscillator method and the balanced-capacitance-bridge methods are suitable. The dynamic method involving measurements of resonance and anti-resonance frequencies, low-frequency capacitance, and resonance resistance is recommended as simplest and most accurate for large-size specimens with high values of Q/r, where r is the ratio of the crystal capacitances. Quasi-static methods are useful for crystals having a low value of Q/r, for determining the sign of the piezoelectric coefficients, and for production checks. Hydrostatic methods are useful for complicated crystals and for measuring a sum of two of the coefficients of electrostrictive ceramics.

621.317.373:621.317.755 2721

Cathode-Ray-Tube Protractor or Synchroscope—H. Sohon. (*Elec. Eng.*, vol. 73, pp. 220–221; March, 1954.) Description of a bridge-type rc phase-shifting network for use with a cro for measurement of the phase difference between two signals. The angle is indicated on the calibrated cr tube screen by the major axis of the elliptical trace.

621.375.3.015.3.001.4 2722

A Transient Analyzer for Magnetic Amplifiers—J. E. Smith. (*Trans. AIEE*, Part I, *Communication and Electronics*, vol. 72, pp. 461–465; 1953. Digest, *Elec. Eng.*, vol. 73, p. 218; March, 1954.) The response time of the magnetic amplifier is obtained directly by comparing the saturation phase angle with a reference phase angle marked by a train of pulses.

621.317.41 2723

An Automatically Recording Magnetic Balance—T. Hirone, S. Maeda and N. Tsuya. (*Sci. Rep. Res. Inst. Tohoku Univ., Ser. A*, vol. 6, pp. 67–76; February, 1954.) Description of instrument for determining the magnetization /temperature characteristics of small samples over the range from the temperature of liquid *He* to $\sim$1000 degrees C.

621.317.411+621.317.335.3:621.318 2724

Theory of Measurement of μ and ϵ of Semiconducting Ferromagnetics—K. M. Polivanov. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 95, pp. 61–64; March 1, 1954. In Russian.] The method used for determining the complex electrical impedances of a specimen in an electric circuit or a magnetic field is analogous to the open- and short-circuit method of determining characteristic impedance and propagation constant of a long line. Expressions for the complex impedances in terms of the complex values of μ and ϵ are given. The advantages of using rod ring or specimens in a coaxial line are noted.

621.317.7:621.396.822:621.397.828 2725

Impulse Noise Generators—M. V. Callendar. (*Electronic Engineering*, vol. 26, pp. 200–203; May, 1954.) Two generators are described, suitable for testing sound or vision receivers for their response to impulsive noise from car ignition systems and other sources. One is based on a spark plug with variable gap and controllable output. The other is a thyratron generator, providing pulses with any desired time relation to the synchronizing pulses. Calibration of these generators is considered.

621.317.7.029.64 2726

Integrated Microwave Test Bench—(*Wireless World*, vol. 60, pp. 351–352; July, 1954.) *X*-band test equipment comprising wavemeter, attenuator, slotted line, etc. is milled out of one solid block of light alloy instead of being assembled from separate units. Losses and undesired reflections are thus reduced.

621.317 7.029.64 2727

Microwave Measuring Equipment—P. M. Ratcliffe. (*Jour. Brit. IRE*, vol. 14, pp. 243–259; June, 1954. Discussion, pp. 260–261.) A survey dealing with instruments for field and factory, as well as for laboratory use, at centimeter wavelengths. The measurable quantities in this band are wavelength (in preference to frequency), power (with emphasis on heating effect), and normalized impedance (in terms of electric and magnetic field). Equipment described includes capacitance-loaded-line and resonant-cavity wavemeters, thermistor-type milliwattmeter, neon-tube tester for swr, directional couplers, spectrum analyzer, noise generator and a complete radar test set for the 3-cm waveband.

621.317.727.029.63 2728

Accurate Radio-Frequency Microvoltages—M. C. Selby. (*Trans. AIEE*, Part I, *Communication and Electronics*, vol. 72, pp. 158–163; 1953.) Coaxial-type rf micropotentiometers developed at the National Bureau of Standards (1712 of 1951) are described in detail.

621.317.734.028.3 2729

An Instrument for Measurement of Very High Resistance—F. J. Lynch and C. L. Wesenberg. (*Rev. Sci. Instr.*, vol. 25, pp. 251–255; March, 1954.) An instrument for accurate measurement in the range 10^9–$10^{13}\Omega$ is described. A constant current flows through the resistor under test and potential drop is measured by an electrometer. The constant current is the displacement current flowing through a standard capacitor when the potential difference across it changes linearly with time. Provision is made for measuring resistance with 3 mv–10v dropped across the resistor. Accuracy of the instrument is within about 0.5 per cent.

621.317.755:621.3.015.3 2730

A Method for the Measurement of Phase and Amplitude Variations of Transient High-Frequency Phenomena—A. Essmann. (*Z. angew. Phys.*, vol. 6, pp. 115–120; March, 1954.) A development is described of the method of Nijenhuis (2485 of 1941) for a continuous visual presentation of phase and amplitude changes produced e.g. by an amplifier, by means of a brightness-modulated spiral cro trace. The complete circuit, operation and application of this phase- and amplitude-analyzer are given and the traces obtained with a two-stage amplifier and with a detuned single-stage amplifier are shown photographically. Phase angles can be read to within ±1 degree.

621.317.78.029.64 2731

40- to 4000-Microwatt Power Meter—R. W. Lange. (*Trans. AIEE*, Part I, *Communication and Electronics*, vol. 72, pp. 492–494; 1953.) Description of an instrument for use at 10 kmc, comprising a temperature-compensated bolometer bridge using a bead thermistor as power detector.

621.397.62.001.4:535.623 2732

Color-Bar Generator produces I-Q Signals—A. F. Boscia. (*Electronics*, vol. 27, p. 143; June, 1954.) Test signals for alignment of the *Q* and *I* demodulators for reception of NTSC color-television signals are obtained from a color-bar generator by control of blue-green and blue-red overlap.

621.317 2733

Electronic Measurements (Book Review)—F. E. Terman and J. M. Pettit. Publishers: McGraw-Hill Book Company, New York, N.Y., 1952, 707 pp., 72s.6d. (*Jour. Atmos. Terr. Phys.*, vol. 5, p. 112; May, 1954.) The emphasis is "on the fundamental principles of measurement and of their synthesis into more elaborate techniques for specific purposes or circumstances. . . . The presentation of the material is comprehensive and yet critical."

OTHER APPLICATIONS OF RADIO AND ELECTRONICS

532.542 2734

Electromagnetic Velometry: Part 2—Elimination of the Effects of Induced Currents in Explorations of the Velocity Distribution in Axially Symmetrical Flow—A. Kolin and F. Reiche. (*Jour. Appl. Phys.*, vol. 25, pp. 409–413; April, 1954.) No correction for the effects of induced currents is required if very fine unshielded wires can be used as electrodes. This is possible if the quantity measured is the potential gradient along a cross-section diameter at 45 degrees to the direction of the magnetic field. Part 1: 3981 of 1945 (Kolin).

615.471 2735

Determination of the Resultant Dipole of the Heart from Measurements on the Body Surface—D. Gabor and C. V. Nelson. (*Jour. Appl. Phys.*, vol. 25, pp. 413–416; April, 1954.) Theory was developed and verified by measurements in electrolyte tank models of the human thorax.

621.372.029.64:621.313.001.4 2736

Microwaves Used to Observe Commutator and Slip Ring Surfaces during Operation—A. H. Ryan and S. D Summers. (*Elec. Eng.*, vol. 73, pp. 251–255; March, 1954.) Variations in the position of a reflecting surface are used to change the effective length of the open collinear arm of a magic-*T*. A change of 0.5 mil can be detected by means of the apparatus described operating at a frequency of 24 kmc. Typical records are shown.

621.384.612 2737

Overcoming the Critical-Energy Effect in the Strong-Focusing Synchrotron—E. Bodenstedt. (*Z. Naturf.*, vol. 8 (a), pp. 502–503; August, 1953.) At a critical energy value the stable equilibrium of the phase oscillation becomes unstable and the focusing action changes into a defocusing one. A theoretical estimation indicates that it is not possible to avoid the transition through this critical value. The difficulty is overcome by using a large even number of accelerating gaps grouped in two systems, and distributing the phase of the accelerating voltages so that particles are focused in passing through the gaps of the first system and defocused in passing through the gaps of the second system; when the particle energy reaches the critical value the two systems reverse.

621.384.622.1 2738

Alternating-Gradient Electrostatic Focusing for Linear Accelerators—L. C. Teng. (*Rev. Sci. Instr.*, vol. 25, pp. 264–268; March, 1954.) Analytical study of the application of alternate convergent and divergent lenses to a standing-wave heavy-particle linear accelerator.

621.385.833 2739

An Electron-Optical Method of Imaging Objects with Magnetic Inhomogeneities—G. V. Spivak, N. G. Kanavina, I. N. Chernyshev, and I. S. Sbitnikova. [*Compt. Rend. Acad. Sci.* (*URSS*), vol. 92, pp. 541–543; September 21, 1953. In Russian.] A secondary-emission electron microscope, by means of which both the geometrical and the magnetic irregularities of the specimen can be detected, is briefly described, typical photographs are shown and the method of operation is briefly compared with the electron-optical shadow method used by Marton and Lachenbruch (1211 of 1950).

621.385.833:061.3 2740

Fifth Annual Convention of the German Association for Electron Microscopy, Innsbruck, 1953—(*Optik* (*Stuttgart*), vol. 11, pp. 97–148; 1954.) The text is given of the following papers presented at the convention:

"Image Formation in the Electron Microscope,"—W. Glaser (pp. 101–117)
"Report of French Work in the Field of Electron Microscopy during the Past Year,"—R. Bernard (pp. 118–120)
"Electron Microscopy at Cambridge,"—V. E. Cosslett (pp. 121–132)
"Electron-Microscope Investigations at the A.E.I. Research Laboratory," Aldermaston, England—M. E. Haine (pp. 133–144)
"Report of Swiss Work in the Field of Electron Microscopy,"—E. W. Schütz (pp. 145–148).

621.387.424 2741

The Temperature Dependence of the Characteristics of (self-quenching) Gas-Filled Geiger-Müller Counter Tubes—H. J. Mader. (*Z. angew. Phys.*, vol. 137, pp. 216–227; March 15, 1954.)

621.387.424 2742

Electron Transit Times in Geiger Counters—J. R. Heirtzler. (*Rev. Sci. Instr.*, vol. 25, pp. 243–245; March, 1954.) Using a photoemission technique, the transit time of an electron between the cylinder and the central wire of a self-quenching Geiger counter has been measured as a function of overvoltage for different vapor fillings at different pressures. Results are presented in graphs.

621.387.424 2743

Particle Detectors of Geiger-Müller (G-M) Counter Type—A. Benoit. (*Le Vide*, vol. 9, pp. 1475–1491; March, 1954.) Operating principles are outlined and technological aspects discussed. 27 references.

621.398 2744

Parachute-Borne Telemetering System—M. L. Greenough and C. C. Gordon. (*Electronics*, vol. 27, pp. 148–151; June, 1954.) Description of a low-cost semi-expendable unit for transmitting data during the parachute descent. Seven information channels are provided, each sampled 100 times per second. The operating range is 2–10 miles. For a shorter account, see *Tech. News Bull.*, *NBS*, vol. 38, pp. 81–82; June, 1954.

621.398+621.396.93]:621.396.665 2745

Sequentially Gated Automatic Gain Control—Eliason. (*See* 2677.)

PROPAGATION OF WAVES

538.566 2746

Some Stochastic Problems in Wave Propagation—J. Feinstein. (*Trans. I.R.E.*, vol. AP-2, pp. 23–30 and 63–70; January and April, 1954.) "The effect of random height variations associated with a conducting surface upon the characteristics of reflected wave energy is ascertained by the methods of physical optics. Average received power, its variance, angular and frequency power spectra, and the field correlation pattern are determined in terms of the statistical parameters of the surface. Volume type problems are treated by ascertaining the effect of refractive index fluctuations within a slab upon an emergent wave-front, and then generalizing to a continuous medium. The results are applied to various problems encountered in tropospheric and ionospheric wave propagation."

538.566:535.312 2747

The Scattering of Electromagnetic Waves by Turbulent Atmospheric Fluctuations—F. Villars and V. F. Weisskopf. (*Phys. Rev.*, vol. 94, pp. 232–240; April 15, 1954.) "The statistical theory of turbulence is applied to the problem of density fluctuations in the troposphere and the ionosphere. For suitable wavelengths, for which the so-called similarity region (Kolmogoroff spectrum) of the spectrum of turbulence is relevant, a closed formula can be given for the scattering cross section. It contains as only parameter the turbulent power dissipation S, and its angular dependence is given by $(\sin\frac{1}{2}\theta)^{13/3}$, θ being the scattering angle. The values of S required to explain ionospheric scattering are in excellent agreement with values found from investigations of meteor trails. Tropospheric data cannot be fitted with the assumptions of dry-air turbulence alone. The inference is that humidity fluctuations play an essential part in tropospheric scattering." If certain assumptions are valid, only a very small value of large-scale humidity fluctuations is required to account for the observed scattering.

621.396.11:551.510.535:523.3 2748

Lunar Radio Echoes and the Faraday Effect in the Ionosphere—W. A. S. Murray and J. K. Hargreaves. [*Nature* (*London*) vol. 173, pp. 944–945; May 15, 1954.] Fading of lunar echoes has been investigated at Jodrell Bank; the frequency used was 120 mc. 50,000 echoes were photographed and analyzed. It is suggested that long-period fading observed during the day is due to rotation of the plane of polarization of the radio waves as they traverse the ionosphere in the presence of the earth's magnetic field; this view is supported by analysis based on the Appleton-Hartree magneto-ionic theory. Confirmation was obtained from experiments in which a vertical and a horizontal receiving dipole were alternately switched into operation.

621.396.11:621.317.353.2:551.510.535 2749

Effect of Nonlinearity of the Medium on Radio Waves Propagated in the Ionosphere—I. M. Vilenski. (*Compt. Rend Acad. Sci.* (*URSS*), vol. 92, pp. 525–528; September 21, 1953. In Russian.) Assuming a square-law nonlinearity of the ionosphere, and neglecting the effect of the geomagnetic field, the field strength of the third harmonic of a monochromatic radio wave is calculated. Substitution of numerical values in the approximate formula derived indicates that the effect should be measurable for a 500-kw transmitter operating at $\omega=10^6$. A more exact calculation is made of the effect produced by a modulated radio wave. The nonlinearity in this case causes not only a change of amplitude and phase but also a noticeable change of the depth of amplitude modulation, production of harmonics of the modulation frequency and the appearance of phase modulation which is large at low modulation frequencies and small at high ones. The expressions, which contain parameters such as the effective number of electron-molecule collisions, may be useful for determining the ionospheric parameters more simply than by the cross-modulation method.

621.396.11.029.53:551.510.535 2750

High Multiple Radio Reflections from the F_2 Layer of the Ionosphere at Brisbane—K. Baird. (*Aust. Jour. Phys.*, vol. 7, pp. 165–175; March, 1954.) Continuous night-time records of multiple F_2 reflections at normal incidence have been made at 2.28 mc, and the echo patterns have been classified. Statistical analyses of occurrences of up to the 10th multiple showed that (a) if no account is taken of the presence or absence of E_s, the frequency of occurrence increases towards dawn, (b) there is no correlation between the number of reflections observed and the virtual height of the region, (c) there is no correlation with "range duplications," (d) inverse correlation between high multiple F reflections and presence of E_s occurs only when the lower region is blanketing, and (e) there is no correlation between high multiples and the traveling disturbances described by Munro (2504 of 1950). High multiples show maxima at the equinoxes. Oblique incidence records indicated no reflections beyond the fifth multiple.

621.396.11.029.6:551.5:061.3 2751

Radio-Meteorology: Conference in Texas—J. A. Saxton. [*Nature* (*London*), vol. 173, pp. 761–764; April 24, 1954.] Report and comment on the conference held in November, 1953. Papers on tropospheric wave propagation dealt with propagation beyond the horizon, scattering by turbulent layers, measurements of atmospheric refractive index, research at 100–1,000 mc by the National Bureau of Standards, prediction of propagation conditions and experiments on 8-mm waves. Other sessions were devoted to thunderstorm and tornado atmospherics, cloud and precipitation physics and weather forecasting from radar observations. See also 230 of January.

621.396.81 2752

The Mechanism and Distribution of Short-Period Fading under Conditions of Ionospheric Turbulence—F. Minozuma and H. Enomoto. (*Proc. Phys. Soc.*, vol. 67, pp. 211–216; March 1, 1954.) "Fading phenomena are discussed in terms of the variations of optical paths produced by the turbulent ionosphere. A simple model of turbulence in the ionosphere can predict the fading speeds found in practice, and some of the physical constants involved can be deduced from the variation of field intensity with time. The technique can be applied to determine fading speed and the amplitude distributions needed for communication purposes"

621.396.812.029.62 2753

V.H.F. Field Intensities in the Diffraction Zone—R. N. Ghose and W. G. Albright. (*Trans. I.R.E.*, vol. AP-2, pp. 35–38; January, 1954.) A solution of the general wave equation is obtained assuming a refractive-index/height variation of exponential form which fits the required boundary conditions. An expression for field strength as a function of the height and the separation of transmitting and receiving antennas is derived. Calculated values of signal strength are compared with measured values.

621.396.812.3.029.62 2754

A Preliminary Study of Fading of 100-Mc/s F.M. Signals—R. L. Riddle and C. R. Ammerman. (*Trans. I.R.E.*, vol. AP-2, pp. 30–34; January, 1954.) A statistical study is made of short-period fading over a mountainous path of about 120 miles under typical midday conditions. Signal amplitudes have an essentially Rayleigh distribution; amplitude variations have a Gaussian distribution. Speed of fading and values of effective wind velocity are determined statistically; the latter are compared with measured values.

RECEPTION

621.396.621 2755

The Response of a Panoramic Receiver to C. W. and Pulse Signals—H. W. Batten, R.A. Jorgensen, A. B. Macnee, and W. W. Peterson. (Proc. I.R.E., vol. 42, pp. 948–956; June, 1954.) A quantitative investigation is made of the dependence of output amplitude, output pulse width and apparent bandwidth of a panoramic superheterodyne receiver on signal frequency, input pulse width, actual bandwidth, sweep rate, and type of IF amplifier. The measured response of a receiver with single-tuned IF amplifier to rectangular-envelope pulses is compared with the theoretical response of a receiver with a Gaussian-shaped IF characteristic to Gaussian-envelope pulses. The agreement between the two cases justifies the application of the Gaussian-case analysis to most practical design problems.

621.396.621:621.396.812.3 2756

An Analysis of Dual Diversity Receiving Systems—A. H. Hausman. (Proc. I.R.E., vol. 42, pp. 944–947; June, 1954.) "A method is presented for evaluating dual diversity receiving systems by relating the characteristics of the receiving equipment to the signal levels at both receiving antennas. The characteristics of the receiving equipment are described by relating the quality of the output traffic to the input signal level. The input signal levels are in turn described in terms of the bivariate Rayleigh probability distribution function. The method is sufficiently general to enable extension to evaluate triple diversity systems and is independent of the frequency range over which diversity fading phenomena occur. The method can also be simplified to examine the over-all effectiveness of a nondiversity receiving system."

621.396.621:621.396.9 2757

Aircraft Receiver for VOR-ILS and Communications—Gray. (*See* 2676.)

621.396.621.029.55:621.396.662.076.2 2758

Notes on Permeability Tuning for Short Waves—P. Rohan. (Proc. I.R.E. (*Australia*), vol. 15, pp. 111–113; May, 1954.) Methods of achieving linear dial calibration are discussed. Formulas are derived for calculating the components of the resonant circuits for band changing and band spreading. Numerical examples are given.

621.396.621.54:621.373.424 2759

Variable-Frequency Crystal-Controlled Receivers and Generators—T. L. Wadley. (*Trans. S. Afr. IEE*, part 2, vol. 45, pp. 77–90; February, 1954. Discussion, pp. 90–99.) The systems described are based on heterodyning the harmonic spectrum of a crystal oscillator by means of a vfo and mixer so that a fixed-frequency filter can be used and interpolation between harmonics is easily effected. General principles applied to generators and receivers are discussed. Detailed consideration is given to the design of (a) a generator for frequencies up to 20 mc derived from a 100-kc crystal, and (b) a receiver operating at frequencies up to 30 mc with multiple heterodyning based on a 1-mc crystal. In the latter the local oscillator is a generator of the type described and the outputs from the vfo and the harmonic filter are applied respectively to the first and second mixers; a final IF of 456 kc is derived in a 2–3-mc interpolation receiver. Alternative circuits are considered and performances discussed.

621.396.621.54.029.5 2760

Wide-Band Communication Receiver—(*Wireless World*, vol. 60, pp. 333–334; July, 1954.) A 12-tube receiver, model CAT, designed to an Admiralty specification for use in naval vessels, has a frequency range of 60 kc to 31 mc, covered in eight switched ranges. The set is a double superheterodyne, but is switched to single-heterodyne operation on some of the ranges. The if bandwidth can be adjusted over a wide range in four discrete steps.

621.396.621.59 2761

A New Method for Treating Electron Tubes when used as Superregenerative Detectors—A. E. Mostafa and M. El-Shishini. (*Trans. AIEE*, Part I, *Communication and Electronics*, vol. 72, pp. 207–213, 283–289, and 290–297; 1953.) Analysis is given based on treating the tube as a periodically varying element and applying the methods of superposition and successive approximation used for the solution of circuits having periodically varying resistances. Multiple resonance, the characteristic noise, the building up of oscillations to an amplitude depending on the quenching source and the tube nonlinearity, and the condition for stability of superregeneration are discussed in relation to the no-signal condition. The avc action, sensitivity, selectivity, stability and synchronization with the signal are considered. Numerical illustrations are given and experimental investigations described.

621.396.823:621.376.2/.3 2762

Impulse Noise in Amplitude-Modulation and Frequency-Modulation Systems—D. Maurice. (*Onde elect.*, vol. 34, pp. 303–307; March 1954.) Results of a B.B.C. survey are discussed. For a fm system with maximum frequency swing of ±75 kc and pre- and deaccentuation with 50-μs time constant, the improvement in signal/noise ratio over a comparable am system with no accentuation is 26 db for both impulse noise and circuit noise above a threshold value of signal intensity. The effect of impulse noise with peak values greater than and less than carrier level on arrival at the fm limiter is analyzed by the rotating-voltage-vector method. A family of curves based partly on calculation and partly on measurement shows how the interference from motor cars in a city street depends on the maximum frequency swing of the fm receiver.

STATIONS AND COMMUNICATION SYSTEMS

621.376.5 2763

Theoretical Fundamentals of Pulse Transmission: Part 1.—E. D. Sunde. (*Bell Sys. Tech. Jour.* vol. 33, pp. 721–788; May, 1954.) Theory is given in a form useful both for investigating existing pulse transmission systems and for designing new ones. The relation between the pulse transmission characteristic and the frequency response characteristic is discussed, and characteristic distortion due to system imperfections is considered. Methods of evaluating pulse distortion from gain and phase deviations are presented, and the resulting limitations on pulse transmission rates in low-pass symmetrical and asymmetrical sideband systems are examined.

621.39 2764

Communications for Civil Defense—C. A. Armstrong. (*Trans. AIEE*, Part I, *Communication and Electronics*, vol. 72, pp. 315–326; 1953.) Discussion of the requirements for dealing with attack warning and disaster control in the U.S.A.

621.391.5:621.395.61 2765

The "Vagabond" Wireless Microphone System—T. W. Phinney. (*Trans. I.R.E.*, vol. AU-2, pp. 44–48; March/April, 1954. *Proc. NEC, (Chicago)*, vol. 9, pp. 103–110; 1953.) Combined microphone and fm transmitter equipment for a public address system, designed to be carried or worn, is described. The system uses inductive coupling between the transmitter and the receiver, and a carrier frequency of approximately 2.1 mc. The subminiature transmitter, which weighs <1 lb. is contained in a stick-type microphone housing $1\frac{1}{4}$ inches in diameter and 12 inches long. Two printed circuit plates form the main chassis, the unit, after assembly, being filled with casting resin. A microphone cartridge of special design provides the desired pre-emphasis. A ferrite-core transmitting inductor 3 inches long is used. Battery life is 40 operating hours.

621.394/.395 2766

Telephony and Telegraphy—L. H. Harris. (*Proc. IEE*, part I, vol. 101, pp. 83–92; May, 1954.) A survey covering developments in materials, components and equipment since 1946. 100 references.

621.396.1(94) 2767

Frequency Bands allotted to Australian Radio Services as at 1/7/53—(*Proc. I.R.E. (Australia)*, vol. 15, pp. 120–121; May, 1954.)

621.396.44 2768

Program Transmission over Type-N Carrier Telephone—R. L. Case and I. Kerney. (*Trans. AIEE*, part I, *Communication and Electronics*, vol. 72, pp. 791–795; 1953.)

621.396.5 2769

The New Jersey Turnpike—A Unique Highway Communication System—P. F. Godey, J. R. Neubauer, and D. R. Marsh. (*Trans. AIEE*, Part I, *Communication and Electronics*, vol. 72, pp. 360–369; 1953.) Description of the radio system for this busy road, comprising seven 950-mc stations forming the backbone of the system, five 150-mc stations for communication with mobile units, about 75 mobile units of the police, maintenance, etc., and about 86 toll booths at 17 interchange stations.

621.396.61/.62 2770

Ships' Lifeboat Radio—(*Wireless World*, vol. 60, p. 308; July, 1954.) A transmitter-receiver approved by the British GPO and Ministry of Transport has an improved kite to raise the antenna to a height of about 200 feet. Daylight ranges of over 500 miles have been obtained with radiated power of 4w. An alternative fixed antenna is supported on an 18-foot mast. Brief details are given of the set, which transmits on 500 kc and 8.634 mc and receives on 500 kc only. Power is provided by a hand-turned generator.

621.396.65.029.6 2771

Microwave Radio Relay Systems Symposium, New York, November 5–6, 1953—(*Trans. I.R.E.*, vol. MTT-2, pp. 1–107, April, 1954; vol. CS-2, pp. 1–107, July, 1954.) The text is given of the following papers presented at the symposium:

"The Microwave System of the Michigan-Wisconsin Pipeline Company,"—W. P. Maginnis and H. Place (pp. 1–8).

"Microwave Site Selection in Undeveloped Country,"—R. D. Pynn (pp. 9–15).

"Microwave Repeater Site Planning and Development,"—R. D. Chapman (pp. 16–31).

"Remote Control of Standby Engine Generator Sets over a Microwave System,"—R. L. Halvorson (pp. 32–35).

"Applications of Compandors to F. M. Radio Systems with Frequency Division Multiplexing,"—M. C. Harp, M. H. Kebby and E. J. Rudisuhle (pp. 36–40).

"A Double-Sideband Amplitude-Modulated Multiplex System for Use over Microwave Radio,"—N. B. Tharp (pp. 41–49).

"A Microwave System for Trunk Service,"—J. J. Lenehan (pp. 50–59).
"A Microwave Radio System for Pipeline Use," (abstract only),—E. Dyke (pp. 60–62).
"Microwave and V.H.F. Radio Installation for the Union Electric System,"—G. W. Fox (pp. 63–83).
"Microwave Radio Relay Link for Military Use,"—S. Metzger (pp. 84–88).
"Transco Microwave System,"—H. A. Rhodes (pp. 89–92).
"Microwave Testing with Millimicrosecond Pulses,"—A. C. Beck (pp. 93–99).
"Theoretical Aspects of Microstrip Waveguides (abstract only),—G. A. Deschamps (pp. 100–102).
"Considerations in Klystron Design for Microwave Relay Systems,"—R. W. Olthuis (pp. 103–107).

621.396.712 2772

The Transmitting Centre of the Institut National Belge de Radiodiffusion (I.R.N.) at Wavre-Overijse—M. Dick. (*Brown Boveri Rev.*, vol. 40, pp. 370–394; October 1953.) Detailed illustrated description of station layout and equipment. See also 1128 of 1953 (Hansen).

621.396.933 2773

The Information Content of Air-Ground Messages—G. W. Grier, Jr. (*Trans. I.R.E.*, vol. ANE-1, pp. 5–16; March, 1954.) Records of air/ground communication during the landing of aircraft were analyzed with a view to assessing (a) the general reliability of the information conveyed, and (b) whether the best possible use was being made of the available channel space. It is suggested that the position- and altitude-reporting functions of the pilot be taken over by a transponder.

621.396.933 2774

Some Channel Allocation Problems in Air-Ground Voice Communications—W. W. Felton. (*Trans. I.R.E.*, vol. ANE-1, pp. 41–51; March, 1954.) Discussion based on measurements of the factors limiting the number of aircraft under simultaneous control per channel, including the maximum utilization of the channel, and the amount of communicating time per aircraft.

621.39.001.11 2775

Communication Theory (Book Review)—W. Jackson (Ed.). Publishers: Butterworths Scientific Publications, London, Eng., 1953, 532 pp., 65s. (*Jour. Atmos. Terr. Phys.*, vol. 5, p. 111; May, 1954.) "The publication of this book constitutes a very satisfactory conclusion of a symposium on 'Applications of Communication Theory,' which was held at the Institution of Electrical Engineers, London in September, 1952. . . . The volume now available comprises the complete text of the thirty-eight papers presented at the various sessions together with the discussions, some of which are most illuminating and considerably enhance the value of the individual contributions."

SUBSIDIARY APPARATUS

621–526:061.3 2776

Servomechanism Papers—(*Trans. I.R.E.*, vol. OT-1, pp. 1–70; March, 1954.) Six papers presented at a convention held in San Francisco, August 1953, with editorial:
"Trends in Feedback Systems,"—O. J. M. Smith (pp. 2–8).
"Nonlinear Control Systems with Random Inputs,"—R. C. Booton, Jr. (pp. 9–18).
"A Method of Analysis and Synthesis of Closed-Loop Servo Systems containing Small Discontinuous Nonlinearities,"—D. T. McRuer and R. G. Halliday (pp. 19–34).
"Stability of Feedback Systems using Dual Nyquist Diagram,"—P. Jones (pp. 35–44).
"Optimum Lead-Controller Synthesis in Feedback-Control Systems,"—L. G. Walters (pp. 45–48).
"On the Comparison of Linear and Nonlinear Servomechanism Response,"—T. M. Stout (pp. 49–55).
"Predictor Servomechanisms,"—L. M. Silva (pp. 56–70).

621.3.013.783 2777

The Use of Steel Sheet for the Construction of Shielded Rooms—A. M. Intrator. (*Trans. AIEE*, Part I, *Communication and Electronics*. vol. 72, pp. 599–604; 1953. Discussion, pp. 604–605.)

621.316.722.1:621.385.3 2778

Self-Heating Triode for Voltage Stabilization—Hopkins. (*See* 2824.)

621.35:537.311.33:535.21 2779

Sun powers Telephone—(*Electronics*, vol. 27, pp. 196, 198; June, 1954.) Brief description of a battery comprising an array of *Si* strips, each about $\frac{1}{2}$ inch×2 inches, which when exposed to sunlight gives a current of $24\ mA/cm^2$ at about 0.5v.

TELEVISION AND PHOTOTELEGRAPHY

621.397.24/.26 2780

European Television—J. T. Dickinson. (*Wireless World*, vol. 60, pp. 319–321; July, 1954.) Discussion of engineering problems involved in the recent international exchange experiments, in which networks operating on 405-, 625-, and 819-line standards were interconnected. The main items of the agreed performance specification are indicated. Results were promising.

621.397.242:621.395.51 2781

The London-Birmingham Television-Cable System—T. Kilvington, F. J. M. Laver, and H. Stanesby. (*Proc. IEE*, part I, vol. 101, pp. 117–119; May, 1954.) Discussion on 2625 of 1952.

621.397.26 2782

Transmission Characteristics of the Berlin-Leipzig Television Link—G. Megla. (*Nachr-Tech.*, vol. 4, pp. 98–102; March, 1954.) The technical and economic considerations which led to the adoption of a radio relay network operating with a frequency of about 1.5 kmc are discussed, and calculations of transmitter powers required for adequate point-to-point service are given. Distances up to 82 km are covered using a 10wfm transmitter with a parabolic-reflector antenna with a gain of 3.63 neper. The effect of atmospheric conditions on field strength is briefly discussed.

621.397.26:621.396.65.029.64 2783

Microwave Relay for Japanese Television—T. Nomura, K. Suzuki, S. Mita, and N. Sawazaki. (*Electronics*, vol. 27, pp. 152–156; June, 1954.) Description of the seven-station, 288-mile Tokyo-Osaka relay, which operates with alternate transmitting and receiving frequencies of 4.000 and 4.045 kmc. The design had to take account of high free-space attenuation and deep fades. Necessary gain was provided by incorporating a 3-stage traveling-wave amplifier giving an output of about 3w at each repeater, and by using high-grain paraboloidal antennas of diameter 13.1 feet. The repeaters are of double-heterodyne type, converting the microwave signal first to 70 mc and then to 115 mc.

621.397.26:621.396.65.029.64 2784

The Manchester-Kirk o'Shotts Television Radio-Relay System—G. Dawson, L. L. Hall, K. G. Hodgson, R. A. Meers, and J. H. H. Merriman. (*Proc. IEE*, part I, vol. 101, pp. 93–109; May, 1954. Discussion, pp. 109–114.) The link comprises seven intermediate stations for a total path length of 247 miles, and operates in the 3.9–4.2 kmc band. Duplicate equipment is provided at each station, with stand-by power supply and automatic changeover facilities. All intermediate stations are unattended, a 4-wire line circuit for supervision and control linking them to the terminal points. Only two carrier frequencies are used, each repeater transmitting on one frequency in both directions and receiving on the other frequency from both directions. Fm is used; the over-all repeater bandwidth is 16 mc. A traveling-wave amplifier provides the output of each repeater. Transmitter power is 1w. Details are given of equipment and propagation tests. Outage time from all causes was <0.04 per cent in the period June 1953–January 1954.

621.397.5 2785

Development of Television Service Standards and Application to Design of a Television Broadcast Network—O. W. B. Reed, Jr. (*Trans. AIEE*, Part I, *Communication and Electronics*, vol. 72, pp. 838–850; 1953.) A discussion of frequency allocation and coverage problems in the U.S.A.

621.397.61 2786

Technical Characteristics of FTL Type No. 20-B U.H.F. Television Transmitter—E. M. Bradburd. (*Trans. AIEE*, Part I, *Communication and Electronics*, vol. 72, pp. 555–561; 1953.) Details are given of both sound and picture transmitters. Fm is used for the sound transmitter and grid modulation in a wide-band amplifier for the picture transmitter. Vestigial-sideband filter and diplexer units are analyzed and the construction of the cavity resonators is described.

621.397.62 2787

Television I. F. Inquiry—G. H. Russell. (*Wireless World*, vol. 60, pp. 322–325; July, 1954.) Review of a report prepared by the European Broadcasting Union on the basis of replies to a questionnaire, addressed to manufacturers in a number of countries, on television receiver problems. The relative importance of six major causes of interference listed varied according to local conditions; in Belgium, Denmark, Germany, Holland, and the U.S.A., the most important consideration was second-channel interference, whereas in Sweden and the U. K. it was if breakthrough. Oscillator radiation from nearby receivers is considered of relatively little importance. Values of if in use and expected to be in use in the near future are tabulated. The different television standards used in Western Europe are summarized, and problems in the production of multistandard receivers are mentioned.

621.397.62:535.623 2788

Color Demodulators for Television Receivers—E. G. Clark and C. H. Phillips. (*Electronics*, vol. 27, pp. 164–166; June, 1954.) A circuit based on use of a split-anode heptode tube is described; this permits high-level demodulation using one tube only. Alternative cascode-connected circuits use double-triode and pentode-triode combinations.

621.397.62.001.4:535.623 2789

Color-Bar Generator produces I-Q Signals—Boscia. (*See* 2732.)

621.397.82 2790

The Measurement of Random Monochrome Video Interference—J. M. Barstow and H. N. Christopher. (*Trans. AIEE*, Part I, *Communication and Electronics*, vol. 72, pp. 735–741; 1953.) Two problems are considered, (a) the relative importance of interference in different parts of the video spectrum, and (b) whether the visual mechanism sums the interference effects in various parts of the spectrum in such a way that the over-all effect can be uniquely related to an over-all measurement. Subjective tests are described; the results indicate that random television interference can be measured with a power meter having frequency weighting, with sufficient accuracy to be of value in the design and maintenance of television transmission circuits.

TRANSMISSION

621.396.61 2791

A Million-Watt Naval Communication Transmitter—J. C. Walter. (*Trans. AIEE*, Part I, *Communication and Electronics*, vol. 72, pp. 369–374; 1953.) See 821 of 1953 (Hobart).

621.396.61:[621.373.421.13+621.375.2 2792

A Survey of Quality Problems in Transmitting Equipment—H. Neck. (*Brown Boveri Rev.*, vol. 40, pp. 395–400; October, 1953.) Tolerances in am transmitters are considered. The operation of a quartz-crystal oscillator for maximum stability is discussed and principles of determining amplifier distortion by intermodulation measurements are illustrated. A method of designing an af amplifier based on equivalent-circuit analysis is outlined; it involves a certain mismatch of components but gives improved response at low and high frequencies at high modulation levels.

621.396.61:621.385–712 2793

Air Cooling of Medium- and High-Power Transmitters—Klein. (*See* 2803.)

TUBES AND THERMIONICS

621.314.632:546.289 2794

The "Photo-Aftereffect" of Germanium Crystal Rectifiers—M. Kikuchi and T. Onishi. [*Jour. Phys. Soc.* (*Japan*) vol. 9, pp. 130–131; January/February, 1954.] The reduction of the rectifier resistance, as a result of illuminating the point contact for 90 seconds, was observed; experiments were made to elucidate the physical processes underlying this phenomenon. The results are briefly discussed.

621.314.632:546.289 2795

Germanium Diodes from Spherical Pellets—W. C. Dunlap, Jr. (*Jour. Appl. Phys.*, vol. 25, pp. 448–451; April, 1954.) 15-mil Ge spheres are produced within a few minutes in very large quantities by blowing molten high-purity Ge from a graphite crucible. The spherical pellets can be annealed, ground, etched, and assembled into diodes by techniques easily adaptable to mass production. The diode has a sphere-plane contact in place of the conventional whisker contact. Peak back-voltages in the range 50–100v are easily obtained.

621.314.7 2796

A Point-Emitter/Junction-Collector Transistor—R. H. Kingston. (*Jour. Appl. Phys.*, vol. 25, pp. 513–515; April, 1954.) A transistor structure using a planar *p-n* junction as a collector, and a point contact as an emitter is analyzed. Theory indicates that for maximum frequency cutoff the plane containing the emitter point should be nearly parallel to the collector junction. The base resistance is a critical function of the spacing between the emitter point and the collector junction. Experimental models with modified emitter collector configuration approaching the *p-n-p* structure give improved frequency cutoff.

621.314.7 2797

The Effect of Junction Shape and Surface Recombination on Transistor Current Gain—A. R. Moore and J. I. Pankove. (PROC. I.R.E., vol. 42, pp. 907–913; June, 1954.) Current gain, α, of alloy-type transistors is calculated for some particular three-dimensional geometries, using a method based on analogy with current flow in a conducting sheet, and assuming that surface recombination is the dominant factor in minority-carrier loss. The hole-flow field is plotted on a scale model. The method enables the surface recombination velocity s to be found from simple measurements; s may be useful as a criterion of surface quality. A calculation indicates that transit-time dispersion, due to unequal path lengths in transistors with nonparallel junctions, is not significant at frequencies below 1 mc.

621.314.7 2798

On the Variation of Junction-Transistor Current-Amplification Factor with Emitter Current—W. M. Webster. (PROC. I.R.E., vol. 42, pp. 914–920; June, 1954.) The modification of the field in the base region by the injected carriers decreases the mean carrier transit time, reducing the effect of surface recombination and leading to an increase of α_{cb} as the emitter current increases. The simultaneous increase of conductivity of the base increases the rate of volume recombination and reduces emitter efficiency. The combination of the two effects gives rise to the observed decrease of α_{cb} at both high and low values of emitter current. The variation of α_{cb} with emitter current is about four times as great in *p-n-p* as in *n-p-n* type transistors.

621.314.7 2799

P-N-I-P and N-P-I-N Junction Transistor Triodes—J. M. Early. (*Bell Sys. Tech. Jour.*, vol. 33, pp. 517–533; May, 1954.) A modified *p-n-p* junction transistor is described in which the *n*-type base region and the *p*-type collector region are separated by a relatively thick region of intrinsic semiconductor almost free of impurity centers. This permits establishment of a thick collector depletion layer at relatively low voltage, thus producing low collector capacitance and high collector breakdown voltage, which in turn lead to an extension of the high-frequency limit of operation and of the permissible power dissipation. Calculations based on theory indicate that oscillations at frequencies as high as 3 kmc may be possible. Experimental units have been produced with emitter diameter 0.01 inches and collector diameter 0.015 inches, having stable gain without compensation of 20.5 db at 10 mc and oscillating at frequencies up to 95 mc. The *n-p-i-n* type transistor is formed by a corresponding modification of the *n-p-n* type.

621.383.27 2800

Investigations of the Dark Current of Secondary—Electron Multipliers with Cs-Sb Pyotocathodes—N. Schaetti, W. Baumgartner and C. Flury. (*Helv. Phys. Acta*, vol. 26, pp. 380–383; June 15, 1953. In German.) Continuation of investigations reported in 1518 of 1953 (Schaetti and Baumgartner). Results for *Cs-Sb* photocathodes are compared with those previously obtained for *Li-Sb*.

621.383.5 2801

Tentative Electronic Interpretation of Inertia Phenomena in Photocells—G. Blet [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 238, pp. 1704–1705; April 26, 1954.] By assuming that the conductivity of the barrier layer is proportional to the number of free electrons per unit volume, curves are derived for the potential build-up on illumination, for various illumination intensities; these agree with experimentally obtained curves. The time constant is a characteristic intrinsic to the unilluminated barrier layer and independent of cell geometry and excitation wavelength.

621.385 2802

How to use Valves—(*Elect. Times*, vol. 125, p. 400; March 18, 1954.) A note summarizing the first two parts of British Standard Code of Practice C.P. 1005, covering general recommendations and receiving tubes, cr tubes, and rectifiers.

621.385–712:621.396.61 2803

Air Cooling of Medium- and High-Power Transmitters—W. Klein. (*Brown Boveri Rev.*, vol. 40, pp. 401–406; October, 1953.) An illustrated description is given of different air-cooling arrangements based on the radiator system described earlier [2933 of 1950 (Schärli)]. For a 135-kw medium-wave transmitter one low-pressure and two high-pressure blowers and an exhaust fan are used. More modern, simplified systems are shown in a 100-kw long-wave and a 25-kw medium-wave installation.

621.385.001.4 2804

A Comparison of 6AK5 and 5654 Tubes—F. A. Paul. (TRANS. I.R.E., No. PGCP-1, pp. 18–32; March, 1954.) Vibration and shock tests are reported. Type-5654 tubes, which are electrically equivalent to Type 6AK5 but are designed for increased reliability, gave very much better results, particularly in vibration tests.

621.385.004.15 2805

Reliability of Electron Tubes in Military Applications—E. R. Jervis. (PROC. I.R.E., vol. 42, pp. 902–906; June, 1954.) A statistical study was made of tubes removed from military equipment as unsatisfactory. The factors leading to the removal are grouped under (a) maintenance and operating conditions, (b) environment, (c) application, and (d) inherent tube weaknesses. About one-third of the tubes were found not to be defective. The results indicate the importance of adequate maintenance routine. Excessively high operating temperature is the predominant cause of tube deterioration. The removal rate of the improved "reliable" tubes is only between a quarter and a third of that for the corresponding prototypes. Further improvement in respect of reliability can be achieved by applying known methods.

621.385.029.6 2806

Miniature Radio Valves—V. Anisimov. [*Radio* (*Moscow*), No. 3, pp. 54–56; March, 1954.] Graphs and tables are given of the characteristics of a new series of Russian 6.3-v subminiature tubes designed for use in usw receivers. The series includes two pentodes, two triodes, a diode, a 150-v voltage stabilizer and a thyratron. The base connections are indicated on the tube label.

621.385.029.6 2807

Space-Charge Waves in Plasma Streams—J. Labus and K. Poschl. (*Arch. elekt. Übertragung*, vol. 8, pp. 49–54; February, 1954.) Extension of analysis given previously [2182 of 1953 (Labus)]. The phase velocities and phase eigenvalues are derived for the case of an electron beam in a finite magnetic focusing field.

621.385.029.6 2808

Application of Perturbation Theory to the Traveling Wave Tubes—M. W. Muller and W. L. Beaver. (*Jour. Appl. Phys.*, vol. 25, pp. 542–543; April, 1954.) An expression is derived for the incremental propagation constant as a function of the interaction term.

621.385.029.6 2809

Amplification of Electromagnetic Waves due to Displacement of an Electron Beam in Crossed Electric and Magnetic Fields bounded by Resistive Walls—O. Doehler and G. Guilbaud. [*Compt. Rend. Acad. Sci.* (*Paris*), vol. 238, pp. 1784–1786; May 3, 1954.] Small-signal analysis is presented for a particular arrangement of the general type described by Birdsall et al. (3147 of 1953). Using values of the field given by Maxwell's equations, a second-degree equation is derived for the propagation constant, the solution of which corresponds to one amplified and one attenuated wave. If one of the walls is perfectly conducting, one of these waves disappears. If the magnetic field exceeds a certain value two further waves are obtained with real propagation constants.

621.385.029.6:621.372.2 2810

Transmission-Line Analog of a Modulated Electron Beam—S. Bloom and R. W. Peter. (*RCA Rev.*, vol. 15, pp. 95–112; March, 1954.) The differential equations expressing wave propagation along an electron beam with laminar flow and along a lossless nonuniform transmission line are shown to be identical within certain limits. Conditions in the beam can hence be investigated by means of established transmission-line theory and by measurements on a transmission-line model of the beam. The method is illustrated by a study of space-charge-wave propagation in a planar diode, using a slotted coaxial-line model with a tapered inner conductor and making measurements of the standing-wave pattern.

621.385.029.6:621.372.2 2811

Space-Charge-Wave Amplification along an Electron Beam by Periodic Change of the Beam Impedance—R. W. Peter, S. Bloom, and J. A. Ruetz. (*RCA Rev.*, vol. 15, pp. 113–120; March, 1954.) The common basis of various types of single-velocity space-charge-wave amplifier is established; the impedance of the beam can be changed by changing the beam voltage or diameter or the diameter of the surrounding drift tube. The transmission-line analogy developed by Bloom and Peter (2810 above) is used to determine the amplification bands of the space-charge wave amplifier by analogy with the filter stop bands of the corresponding line.

621.385.029.6:621.373.423 2812

Analysis of the Backward-Wave Travelling-Wave Tube—H. Heffner. (Proc. I.R.E., vol. 42, pp. 930–937; June, 1954.) For low beam currents the tube acts as a high-gain high-Q voltage-tuned filter. As the beam current is increased, self-oscillation is produced, the frequency of which is controlled by the beam voltage to any value within the pass band. Analysis is presented for small-signal operation. See also 587 of February.

621.385.029.6.002.2 2813

Manufacture of a Magnetron—(*Elect. Commun.*, vol. 31, pp. 2–28; March, 1954.) Photographs illustrate various stages in the manufacture and testing of the Type-4J52 magnetron.

621.385.032.213.1.027.5/.6 2814

High-Power Industrial Vacuum Tubes having Thoriated-Tungsten Filaments—R. B. Ayer. (*Trans. AIEE*, Part I, *Communication and Electronics*, vol. 27, pp. 121–125; 1953.) Tubes of this type (2660 of 1952) have given uninterrupted service for periods up to 22 months at anode voltages up to 17 kv.

621.385.032.216 2815

Negative Ion Emission from Oxide-Coated Cathodes: Part 1—F. A. Vick and C. A. Walley. (*Proc. Phys. Soc.*, vol. 67, pp. 169–176; March, 1954.) Simple mass-spectrometer investigations are described. The predominant ions were Cl, originating mainly in the glass envelope. A correlation between Cl ion current and cathode poisoning was established, the poisoning being appreciable when the modulator was more than 8v positive with respect to the cathode.

621.385.032.216 2816

Negative Ion Emission from Oxide-Coated Cathodes: Part 2—W. Grattidge and A. A. Shepherd. (*Proc. Phys. Soc.*, vol. 67, pp. 177–186; March 1, 1954.) Report on mass-spectrometer investigation of the ions emitted on activation and during cathode life. The variation of Cl ion emission with temperature and field was studied. No correlation with electron emission was found. It is probable that Cl ions are mainly responsible for ion burn in cr tubes.

621.385.032.216 2817

The Activation of Thermionic Cathodes Coated with Thoria and Zirconia—G. Mesnard. (*Jour. Phys. Rad.*, vol. 15, pp. 151–155; March, 1954.) Experiments using ZrO_2 on W and Mo bases, ThO_2 on Pt and Mo bases, and complex layers give results suggesting that dissociation of the oxide may be an important cause of activation. For a more detailed account of the work see *Le Vide*, vol. 9, pp. 1448–1453; January, 1954.

621.385.032.216 2818

Moulded Thermionic Cathodes made of Nickel and Alkaline Earth Oxides—G. Mesnard and R. Uzan. (*Le Vide*, vol. 9, pp. 1492–1507; March, 1954.) The method of preparing cathodes by compressing and sintering a mixture of Ni powder and co-precipitated barium and strontium carbonates is described. Measurements on experimental tubes are reported, for steady-state and for pulsed operation. The effects of current activation and heat treatment are demonstrated. The influence of sintering temperature and atmosphere and of the composition of the mixture is investigated.

621.385.032.216 2819

On the Electron Bombardment Effect for Deposited Barium Oxide Films—T. Imai. [*Jour. Phys. Soc.* (*Japan*), vol. 9, pp. 28–37; January/February, 1954.] In tubes with oxide cathodes, oxide films are deposited on the other electrodes by evaporation during pumping, activation, and normal operation. When these films are subjected to electron bombardment, as in the case of anode and screen grid, gas is liberated, and the cathode emission decreases in consequence. The emission from the films increases or decreases on electron bombardment, according as the films are deposited at low or high cathode temperature.

621.385.032.216:[546.431-31+546.431.64 2820

Electronmicroscopic Observation of the Change of Forms of the Coated Oxide Particles of the Oxide Cathode by the Activation—T. Hibi. (*Sci. Rep. Res. Inst. Tohoku Univ.*, *Ser. A*, vol. 5, pp. 573–580; December, 1953.) Results indicate that the degree of activity of oxide particles is related to structural changes occurring at various depths from the surface, as shown in 16 photographs. The changes in barium carbonate particles are also shown photographically and are briefly discussed.

621.385.15 2821

Potential Distribution and Prevention of a Space-Charge-Induced Minimum between a Plane Secondary Electron Emitter and Parallel Control Grid—G. C. Sponsler. (*Jour. Appl. Phys.*, vol. 25, pp. 282–287; March, 1954.) The critical electrode spacing required to avoid formation of a minimum is derived from a potential-distribution analysis.

621.385.15+621.385.832]:621.3.018.75 2822

Some Experiments on Beam-Deflection Devices and Electron Multipliers for Millimicrosecond Pulses—V. H. Vincent. (*Onde elect.*, vol. 34, pp. 119–122; February, 1954.) "An experimental tube with two fast deflector systems has been built and used to indicate coincidences between pulses in coaxial lines with millimicrosecond accuracy. Another tube incorporating an electron multiplier is being constructed as a possible means of amplifying millimicrosecond pulses and for the purpose of studying the characteristics of electron multipliers when delivering large current pulses of millimicrosecond duration."

621.385.15.032.23 2823

Methods of Processing Silver-Magnesium Secondary Emitters for Electron Tubes—P. Rappaport. (*Jour. Appl. Phys.*, vol. 25, pp. 288–292; March, 1954.) Processing experiments under controlled conditions involving various atmospheres, baking times and temperatures were carried out. To produce the best emitter, in which subsequent evaporation of Mg after overheating was almost eliminated, the Ag-Mg alloy was baked first in a low-pressure water-vapor atmosphere for 30 minutes at 550 degrees C., then at standard pressure in an oxygen atmosphere for 5 minutes at 700 degrees C.

621.385.3:621.316.722.1 2824

Self-Heating Triode for Voltage Stabilization—E. G. Hopkins. (*Wireless Eng.*, vol. 31, pp. 169–171; July, 1954.) The tube described has two planar oxide cathodes with a planar grid between them, and two auxiliary anodes which clamp the grid circuit alternately to the two cathodes. A circuit is shown in which the tube is connected between an ac supply and a small fixed load, and acts as a stabilizing variable resistance.

621.385.3/.4].001.4:621.396.822.029.45 2825

Oscillographic Measurements of Valve Noise in Audio-Frequency Channels—G. V. Subhadramma. (*Indian Jour. Phys.*, vol. 27, pp. 359–367; July, 1953.) Measurements made on one tetrode and several triode tubes over the frequency range 375–1320 cps are reported. For a given frequency channel, the triode curves for noise/filament-current exhibit a maximum, in some cases followed by a minimum, for a given anode voltage, while the tetrode curve exhibits a saturation effect. The position of the maximum is the same for all frequencies, but the height of the maximum increases with frequency. For given filament current the amount of noise varies inversely as anode voltage. The absolute values of equivalent noise voltage range from about 20 to 200 μv.

621.385.3.029.62 2826

V.H.F. High-Power Transmitting Tubes—R. Hubner. (*Brown Boveri Rev.*, vol. 40, pp. 417–421; October, 1953.) Description of the construction of BTL-type triodes with thoriated tungsten cathodes and air cooling.

621.385.832:681.142 2827

An Electron-Beam Tube for Analog Multiplication—Angelo. (*See* 2583.)

621.387:621.375.2.024 2828

Transient Response of Glow Discharges with Applications—R. S. Mackay and H. D. Morris. (Proc. I.R.E., vol. 42, pp. 961–964; June, 1954.) Experiments were made with neon and argon lamps, the discharge being maintained by a fixed voltage with a signal voltage superimposed; the voltage variation across a series resistance was observed oscillographically. Transient responses consistent with observed frequency responses were obtained, the response time and characteristic overshoot corresponding to an inductance approaching 1H. Application of standard neon lamps as level-changing coupling elements in direct-coupled amplifiers is indicated.

Collins' new research and development facility in Cedar Rapids.

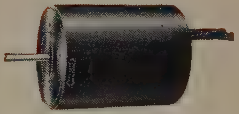
Dallas, Texas, facilities.

Burbank facilities.

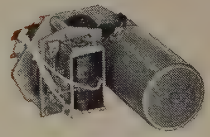
From COLLINS *outstanding research and development facilities* come these contributions to industry

Collins' new research and development facility in Cedar Rapids (pictured top above) houses 12 ultra modern laboratories, with especially designed equipment required by Collins in its diversified development program. It brings together, under one roof, highly trained engineering teams and the most advanced electronic research tools. This new Collins facility in Cedar Rapids is supplemented by Dallas and Burbank laboratories for the development of specialized electronic components for industry.

In addition to the finest facilities, Collins places even greater importance on engineering personnel. Some of the best engineering talent in the nation is at work on Collins' development projects. So today, with greatly expanded research and development laboratories, Collins is better equipped than ever to assist the Electronic Industry. Technical literature is now available on the Collins components shown, and your inquiry is invited.

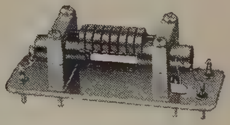


Hystersis Motors



Autotunes and Autopositioners

Oscillators



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MODEL 700
places and measures separate voltage on each element of a tube.

MODEL 7001 Null Reading Device:
Provides tube test accuracy of 1%.

TESTS TUBES PER MANUFACTURERS MANUALS AND JAN SPECIFICATIONS

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This new instrument places a separate voltage on each element of the tube. These voltages can be varied and measured by means of separate variable rheostats and meters in each circuit. AC ripple has been completely filtered out of the plate, screen and grid circuits.

Invaluable for development and research work in studying the behavior of various tubes when used in non-conventional and special circuits.

- Built with HICKOK Highest Quality hand calibrated electrical indicating meters.
- Micromho Ranges: 600, 1500, 3000, 6000, 15,000, 30,000 and 60,000.
- Four Separate Signal Voltages: 1.0, 0.5, 0.1 and 0.05 volts.
- This HICKOK Laboratory Tube Tester is supplied complete and consists of power supply, tube sockets, all necessary leads and accessory material. No additional equipment is necessary.

Write today . . . *for complete technical details on this most complete and accurate tube tester.*

THE HICKOK ELECTRICAL INSTRUMENT CO.
10551 Dupont Avenue • Cleveland 8, Ohio

IRE People

(Continued from page 78A)

Raymond E. Tarpley (A'30), a supervisory research technologist with Leeds & Northrup Company, recently observed the twenty-fifth anniversary of his employment by the instrument manufacturer.

He joined L&N in 1929 after receiving his Master of Science degree from the University of Illinois.

Mr. Tarpley holds a number of patents in the field of woven-type resistors for electrical measurements, and is joint author of a dozen or more papers for scientific magazines. He has served on several committees of the Leeds & Northrup Cooperative Association, and is a member of the AIEE and Franklin Institute.

◆

Lloyd Arthur Ottenberg (S'49–A'51) recently has joined the Kahn Research Laboratories in Freeport, New York, as a Research Engineer. His duties include the development of television and single-sideband equipment.

In 1951, he received the B.E.E. degree from the College of Engineering, New York University.

Before joining Kahn Research Laboratories, Mr. Ottenberg was a member of the engineering staff of Radiomarine Corporation of America, a division of RCA.

◆

W. Evan-Jones (A'43), Ontario Sales Supervisor with the Canadian Marconi Company, Toronto, Ontario, has been named Assistant Sales Supervisor of Commercial Radio Communication Equipment, and transferred to Head Office, Montreal. He will be supervising the sales engineering and sales promotion work of the company's district offices from coast to coast.

Prior to joining the Canadian Marconi Company last year, Mr. Evan-Jones was with the Canadian Comstock Company Limited, Frequency Conversion Division, St. Catharines, Ontario; the Communications Group of the Hydro Electric Power Commission of Ontario; and Northern Electric Company Limited.

He has been actively engaged in the communication field for the past 12 years and is an associate member of the American Institute of Electrical Engineers.

◆

Floyd A. Paul (S'46–A'50–M'52) until recently with the Missile Guidance Department at Northrop Aircraft, has joined the Pacific Division of Bendix Development Laboratories in Burbank, California. He is organizing an Electronic Reliability group concerned with Component Part activity as well as specifications and environmental testing.

(Continued on page 84A)

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Permits examination of any signal under observation without the amplitude loss or waveform distortion generally experienced with conventional R-C probes. The low input capacitance (less than 2 MMF) results in a minimum of loading, particularly on high impedance circuits. Invaluable for operational checks on computers, TV sync amplifiers, and pulse circuits.

FEATURES: **MODEL HF-3**

- Less than 2 MMF input capacitance.
- Input Impedance 4.5 megohms.
- Output Impedance 66 ohms.
- Band Width 10 cps to 10 mc.
- Regulated supply, D.C. on filament.
- 0.25 MV noise level.

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LINE AMPLIFIER

Specifically designed for the transmission and distribution of color television signals in the studio, station or laboratory. Supplies several times the output of conventional video line amplifiers without compression or gamma distortion. Invaluable for piping color signals around the factory. Used by leading color TV manufacturers.

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- Overall Gain 0 to 5X, continuously variable.
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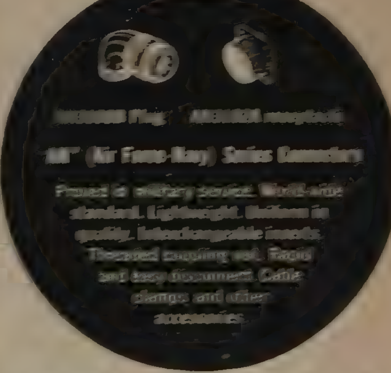

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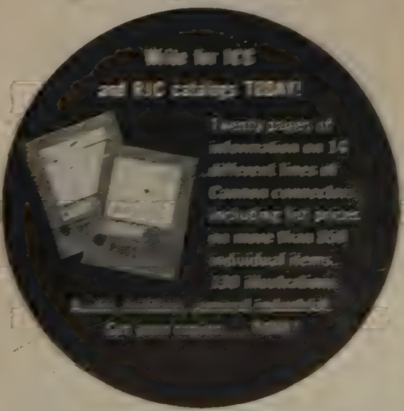

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Twenty pages of information on 14 different lines of Cannon connectors, including list prices on more than 350 individual items. 330 illustrations.

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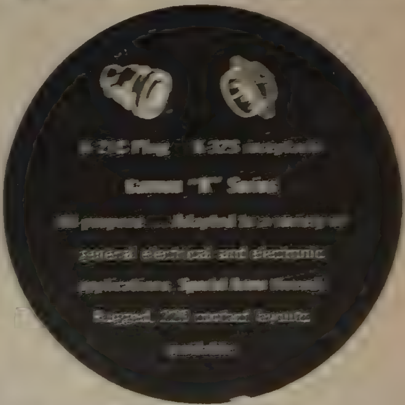

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FREQUENCY RANGE:
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Any value from .1 db to 60 db
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50 OHMS

IRE People

(Continued from page 82A)

Mr. Paul received the B.E. degree in electrical engineering in 1951 at the University of Southern California. For two years, he was a research engineer in the field of component parts and reliability at the California Institute of Technology Jet Propulsion Laboratory.

"Characteristics of Deposited Carbon Resistors" and "Commercial, Ruggedized, and Premium Vacuum Tubes" are two of a number of technical articles he has published. Mr. Paul was Secretary of the Final Steering Committee of the 1953 Electronic Components Symposium.

❖

Chen To Tai (S'44–A'47–SM'50) has recently been appointed Associate Professor in the Department of Electrical Engineering at Ohio State University, where he will teach antenna courses and do research in the Antenna Laboratory. Dr. Tai is well known for his many papers on antenna theory.

Dr. Tai was born in 1915 at Soochow, China. He came to the United States in 1943 and received his D.Sc. degree from Harvard University in 1947. From 1943 to 1949, he was a teaching fellow and a research associate at Harvard University, specializing in antennas. In 1949 he went with the Stanford Research Institute where he was Senior Research Physicist.

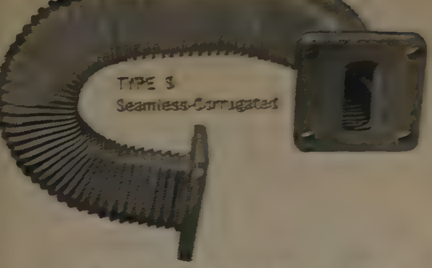
Dr. Tai is a member of Sigma Xi and the American Physical Society.

❖

E. Finlay Carter (A'23–F'36), vice-president and technical director of Sylvania Electric Products Inc., New York, as of October 1 has accepted the position of manager of research operations of Stanford Research Institute. He will supervise the three research divisions carrying SRI's main programs and projects—Physical Sciences, Engineering, and Economics. Mr. Carter will retain in association with Sylvania, assisting in contacts with the Department of Defense and consulting on major research and development contracts.

A native Texan, Mr. Carter was graduated with a degree in electrical engineering from Rice Institute in 1922. Shortly thereafter he joined the General Electric Company as a radio development engineer. In 1929 he became director of the engineering division of United Research Corp., N. Y., a subsidiary of Warner Bros. He went to Sylvania in 1932 as a consulting engineer, later becoming radio division engineer at Emporium, Pa., then assistant chief engineer. In 1941 he was appointed Director of their new Industrial Relations Department, subsequently being elected a vice-president of the company. In 1946 he became vice-president in charge of engineering, and in 1953 vice-president and technical director.

He is a member of Tau Beta Pi, the AIEE, and the Illuminating Engineering Society.



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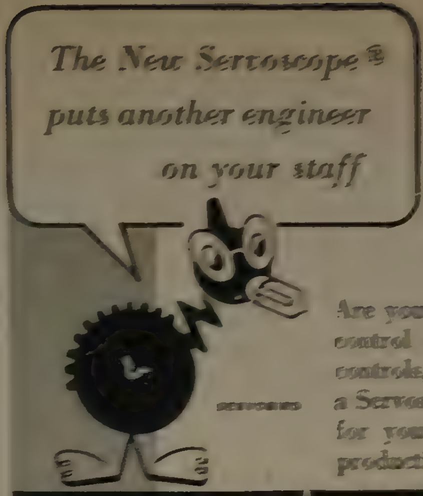

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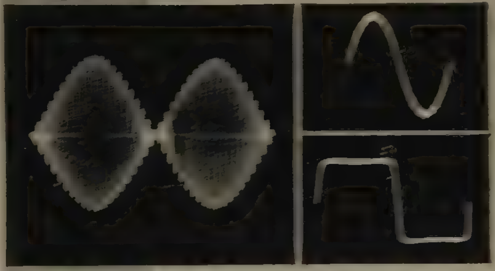
Are you engineering any type of feedback control system—regulators, governors, process controls, positioning or speed servos? With a Servoscope an extra engineer will be working for you on design synthesis, analysis or production test. • Breadboard your intended servo system or other circuit designs—then, by either the frequency response or the transient response method, magnitude and phase curves can be obtained directly within minutes. • For detailed information on how this versatile test instrument can save manpower, materials and money, write Dept. IRE-1



Output wave forms of Servoscope displayed against internal linear sweep generator, frequency is [illegible]

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Have a special problem? Write, wire or phone for details. TODAY! Catalog available.

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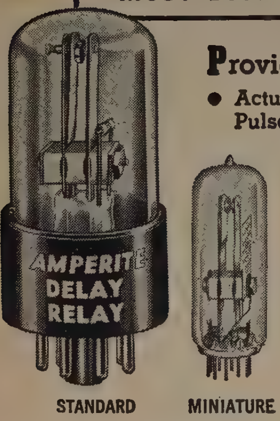
AMPERITE THERMOSTATIC DELAY RELAYS

MOST ECONOMICAL, HERMETICALLY SEALED



Provide delays ranging from 2 to 120 seconds.

- Actuated by a heater, they operate on A.C., D.C., or Pulsating Current.
  - *Hermetically sealed.* Not affected by altitude, moisture, or other climate changes.
  - Circuits: *SPST only*—normally open or normally closed.

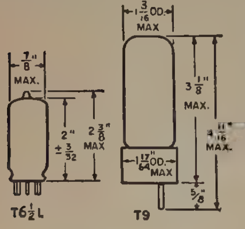
Amperite Thermostatic Delay Relays are compensated for ambient temperature changes from —55° to +70°C. Heaters consume approximately 2 W. and may be operated continuously. The units are *most compact, rugged, explosion-proof, long-lived,* and — *inexpensive!*

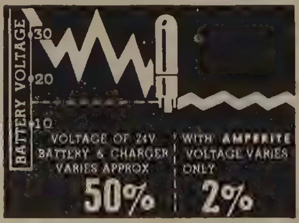
TYPES: Standard Radio Octal, and 9-Pin Miniature.

PROBLEM? Send for Bulletin No. TR-81

BALLAST-REGULATORS

- Amperite Regulators are designed to keep the current in a circuit ***automatically regulated*** at a definite value (for example, 0.5 amp).
- For currents of 60 ma. to 5 amps. *Operates on A.C., D.C., Pulsating Current.*
- Hermetically sealed, light, compact, and most inexpensive.





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Maximum Wattage Dissipation: T6½L—5W. T9—10W.

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Amperite Regulators are the simplest, most effective method for obtaining ***automatic regulation*** of current or voltage. ***Hermetically sealed,*** they are not affected by changes in altitude, ambient temperature (—55° to +90°C), or humidity. Rugged; no moving parts; changed as easily as a radio tube.

Write for 4-page Technical Bulletin No. AB-51

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In Canada: Atlas Radio Corp., Ltd., 560 King St. W., Toronto 2B

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 76A)

Bliley Appoints Kent

Roscoe Kent, 495 43 Ave. N., St. Petersburg, Fla., has just been appointed to represent the **Bliley Electric Co.,** Eire, Pa.

Kent will cover the state of Florida with the Bliley line of quartz crystals, solid ultrasonic delay lines, crystal ovens, and frequency standards.

Transformer Catalog

Electran Mfg. Co., 1901 Clybourn Ave., Chicago 14, Ill., has a new catalog describing their line of transformers, saturable reactors, chokes, special windings, and electronic devices.

These components are custom engineered and built, hermetically sealed, encapsulated, open style, oil filled, and vacuum varnished or waxed.

(Continued on page 107A)

PROFESSIONAL GROUP MEETINGS

Engineering Management

The Philadelphia Chapter, with Mr. A. Emurian presiding, met on June 15 at the Franklin Institute. Mr. Ralph I. Cole of Melpar, Incorporated spoke on "The Influence of Environment on Engineering Careers." Among the points brought out by Mr. Cole were that loyalty ties should be provided in both directions and that engineering staffs should be sounded on what they do not like as well as what they do. A lively discussion period followed the presentation.

For the coming year, the following chapter officers were elected; Mr. A. H. Kettler, Chairman; Mr. J. R. Weiner, Vice Chairman; Mr. G. J. Laurent, Secretary.

(Continued on page 88A)

only Sprague makes them all!

YOU CAN CHOOSE FROM 5 DIFFERENT STYLES OF TANTALEX* CAPACITORS

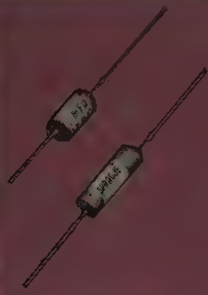
Looking for tantalum electrolytic capacitors? You'll save time and trouble by checking Sprague's complete selection *first*. Sprague makes more types of tantalum capacitors than *any other manufacturer*.

Sprague Tantalex capacitors provide maximum capacitance in minimum space . . . exhibit no shelf aging under long testing periods . . . have extremely low leakage current. And most important, they give unusually *stable* performance, because they're made with tantalum, the most stable of all anodic film-forming materials.

There's a complete range of sizes and ratings available in Tantalex capacitors . . . from the ultra-miniature 10 mf, 4 volt unit in a case only ⅛" in diameter by 5⁄16" long . . . to the 7 mf, 630 volt unit in a case 1⅛" in diameter by 2 19⁄32" long. As for case styles, Sprague makes them all, from tiny tubular and cup units to the large cylindrical types.

For complete details relating to your miniaturization or high temperature problems, write Sprague Electric Co., 233 Marshall St., North Adams, Mass.

Sprague, on request, will provide you with complete application engineering service for optimum results in the use of tantalum capacitors.

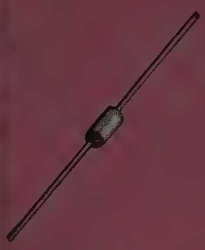


NEW! TYPE 101D for low-cost transistor circuitry

Especially useful for filter, coupling, and bypass applications in transistor electronics, these foil type miniature Tantalex capacitors were intended for use in hearing aids, pocket radios, and similar uses. Operating temperature range is —20 to +65°C. Request Engineering Bulletin 353.

NEW! TYPE 102D for −55°C to +85°C operation for military use

Here are tubular capacitors hermetically sealed in cases of silver plated copper. Intended for applications from 3 to 150 vdc, their small capacitance drop-off at extremely low temperatures, extremely low leakage current, and low power factor are of particular interest. Request Engineering Bulletin 351.



NEW! TYPE 103D ultra-miniature capacitors for transistor circuitry

Only ⅛" in diameter, and from 5⁄16" to ½" in length, these are the smallest electrolytics made. Providing relatively large values of capacitance in the very minimum of space in bypass, coupling, and filter applications, they are ideally suited for transistor hearing aids and military amplifiers in which small size is all-important.
Request Engineering Bulletin 352.

NEW! TYPE 104D miniature "cup" capacitor for military use

These low-voltage units consist of a sintered porous tantalum anode housed in a miniature silver thimble, which serves as both cathode and container for the electrolyte. Volume is less than 1/10 cubic inch; operating temperature range —55 to +85°C, and up to 100°C with a voltage derating of 15%. Request Engineering Bulletin 354.

TYPE 100D for −55 to +125°C operation for military use

These hermetically sealed capacitors are available in voltage ratings up to 630 volts at 85°C or 560 volts at 125°C. They are of the sintered porous tantalum anode type, with internal construction to withstand high g shock, severe vibration, and thermal cycling. Request Engineering Bulletin 350A.

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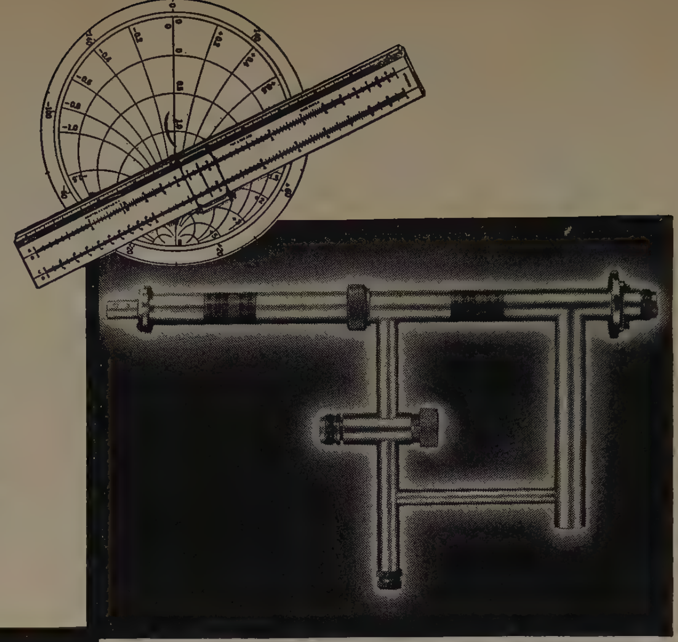

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PROFESSIONAL GROUP MEETINGS

(Continued from page 86A)

CIRCUIT THEORY

The Seattle Chapter held a meeting on June 10 at Bagley Hall, University of Washington. The subject of transistors was presented from three different aspects. "The Physics of Transistors" was the name of the talk made by Mr. S. C. Lawrence of the Physical Research Unit, Boeing Airplane Company. Mr. R. L. Wallace, Jr., of Bell Telephone Laboratories, spoke on "A Junction Transistor Tetrode for High Frequency Use." The last presentation, "Subminiature Transistor Circuits," was made jointly by Mr. R. L. Campbell and Mr. Kermit Thompson, both of the Physical Research Unit, Boeing Airplane Company.

COMPONENT PARTS

The Washington, D. C. Chapter of the Professional Group on Component Parts met on June 23 at the National Bureau of Standards Lecture Room. Mr. Max M. Tall of the Vitro Corporation of America spoke to the group on "Component Specifications; Their Significance and Limitations." Presiding was Mr. Joseph Kaufman, Vice-Chairman.

INFORMATION THEORY

The Los Angeles Chapter held a meeting at Truman's Restaurant on June 10 with Chairman R. R. Bennett presiding. "After Information Theory" was the subject on which Dr. Simon Ramo, of the Ramo-Wooldridge Corporation, spoke. Dr. Ramo suggested that within the next ten years some of the limitations now restricting information theory will be removed by fundamental developments, and numerous applications, especially in the field of industrial management, will be found.

MICROWAVE THEORY AND TECHNIQUES

The Albuquerque Chapter of the Professional Group on Microwave Theory and Techniques held a combined business and technical meeting on June 16, 1954.

Mr. Ted Church led a discussion on "The Theory and Applications of Microstrip." The discussion was centered primarily on the mechanical and electrical characteristics of commercially available components.

For the coming year the membership elected the following officers: Burt J. Bittner, Chairman; Merrill C. Jones, Vice-Chairman; Sheldon Dike, Secretary-Treasurer.

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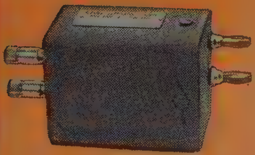

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SECTION MEETINGS

ALBUQUERQUE-LOS ALAMOS

Installation of officers; June 11, 1954.

"A New Wide Band Cathode Ray Oscilloscope," by M. G. Scheraga, DuMont Laboratories; August 16, 1954.

EL PASO

"RCA Color TV," by J. W. Wentworth, Radio Corp. of America; July 30, 1954.

HUNTSVILLE

"Present Development of the Transistor," by T. H. Weaver, Redstone Arsenal; June 29, 1954.

"Phonetics and Communications—Some Engineering Aspects of Our Language," by Dr. B. J. Dasher, Georgia Institute of Technology; July 30, 1954

INYOKERN

"A High Power Magnetic Modulator," by E. R Ingersoll, North American Aviation Co.; August 9, 1954

ROME-UTICA

"Plans and Programs," business meeting; July 27, 1954.

ST. LOUIS

"High Fidelity Sound—Past and Future," with demonstration, by Marvin Camras, Armour Research Foundation; May 27, 1954.

"Horizontal and Vertical Stability as Applied to the Radio Engineer," with demonstration and practical application" election of officers; June 15, 1954

SUBSECTIONS

ERIE

"Overtone Crystal Oscillator Circuitry," by G. H. Lister, Electronic Research and Development Company; January 21, 1954.

"Color Television," by Karl Wendt, Wendt & Squires, Inc.; April 19, 1954.

USAFIT

Discussion and demonstration on reliability of aircraft systems and use of (a) computer, (b) digital autopilot developments, (c) magnetic amplifiers, (d) transistors, (e) system tie-in of design and development of electronic equipment and (f) digital information transfer to operational characteristics; July 30, 1954.

STUDENT BRANCH MEETINGS

UNIVERSITY OF CINCINNATI (IRE-AIEE BRANCH)

General meeting; June 22, 1954.
General meeting; July 19, 1954.

DREXEL INSTITUTE OF TECHNOLOGY (IRE-AIEE BRANCH)

"Applications of Transducers to Electronic Equipment," by Roy Bonsall, Brown Instrument Company; July 15, 1954.

"An Insight into the Career Opportunities of Engineers," by Mr. Batchelder, General Electric Company; August 19, 1954.

ILLINOIS INSTITUTE OF TECHNOLOGY (IRE BRANCH)

"Elements of Transistor Circuits," by R. W. Bull, Armour Research Foundation; May 19, 1954.

NORTH CAROLINA STATE COLLEGE (IRE BRANCH)

Business meeting and installation of officers; May 19, 1954.

MEMBERSHIP

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Cummings, C. I., 1205 Medford Rd. Pasadena 8, Calif.
Davis, V. M., 1197 Hilltop Rd., Kansas City 4, Kan.
Doeleman H., 421 N. Taylor Ave., Montebello, Calif.
Dow, O. E., 110 Hellet Ave., Port Jefferson, L. I., N. Y.
Ebert, I. O., Electrical Engineering Department, Michigan State College, East Lansing, Mich.
Ford, J. P., 392 Harvard St., Oxnard, Calif.
Ganzenhuber, J. H., 3930 Los Feliz Blvd., Los Angeles 27, Calif.
Greenwood, H. T., III, 310 Stanwick Rd., Moorestown, N. J.
Harmon, W. R., International Broadcasting Service, APO 74, c/o Postmaster, San Francisco, Calif.
Heim, H. J., 206 Dehart St., W. Lafayette, Ind.
Kahn, L. R., 22 Pine St., Freeport, L. I., N. Y.
King, L. H., Avien-Knickerbocker, Inc., 58-15 Northern Blvd., Woodside 77, L. I., N. Y.
Lindenberg, E. C., 118 Edison Ave., Erlton, N. J.
Matthews, A. R., Box 3400, USAFIT, Wright-Patterson AFB, Ohio
Morita, T., Stanford Research Institute, Stanford, Calif.
Mullins, F. G., Jr., 318 Oaklee Village, Baltimore 29, Md.
Owens, H. L., 1913 Marconi Rd., Belmar, N. J.
Peters, J. L., 114 Dikeman St., Hempstead, L. I., N. Y.
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Rothenberg, S. M., 120-24—222 St., St. Albans 11, L. I., N. Y.
Thompson, M. F., 9230 Van Ruiten, Bellflower, Calif.
Tomash, E., 2601 Wilshire Blvd., Los Angeles 5, Calif.
Urkowitz, H., 1361 Wells St., Philadelphia 11, Pa.
Weinstock, W. W., 8500 Provident Rd., Philadelphia 19, Pa.
White, C. E., 818 N. Harrison St., Arlington 5, Va.
Wilson, G. L., 198 Bold St., Hamilton, Ont., Canada
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Zarem, A. M., 2325 E. Monte Vista, Pasadena 8, Calif.

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Bondley, R. J., 522 Sanders Ave., Scotia 2, N. Y.
Boykin, E. M., 651 El Medio Ave., Pacific Palisades, Calif.
Buck, L. G., Bell Telephone Company of Canada, Box 6074, Montreal, Que., Canada
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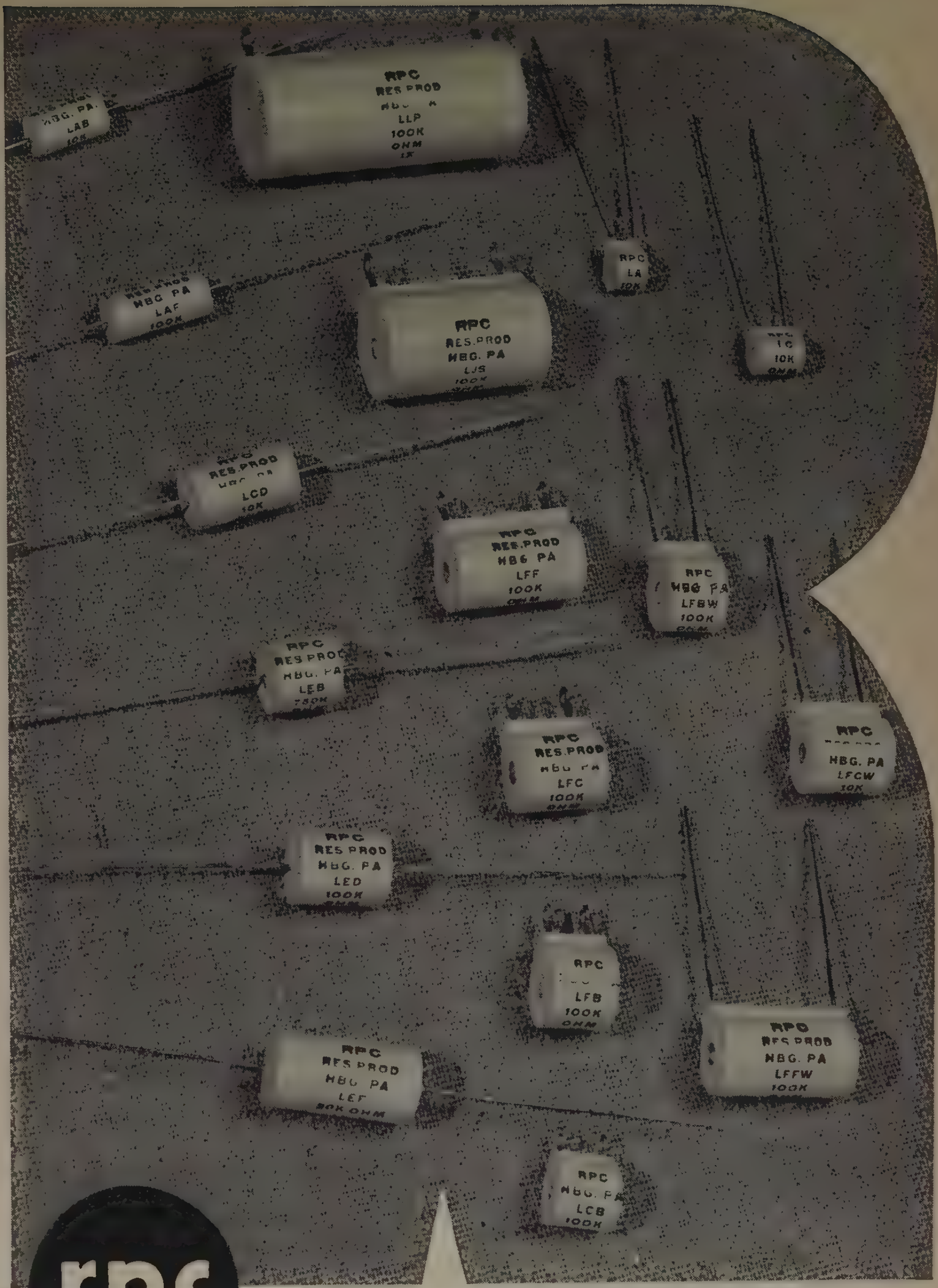
Courtillot, E. P., 88 Rue Raynouard, Paris 16e, France
Heacock, D. P., RCA Victor Division, Harrison, N. J.
Kleinsteuber, W. G., 78, Sierenmoosstr., Konstanz, Germany
McCallen, P. M., Major USAF, 504 S. Walnut St., O'Fallon, Ill.

(Continued on page 95A)



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Blacksmith, P., Jr., Box 3307, USAFIT, Wright-Patterson AFB, Ohio
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Flock, W. L., Department of Engineering, University of California, Los Angeles 24, Calif.
Futterman, N., 1241 W. Boulevard, Los Angeles 19, Calif.
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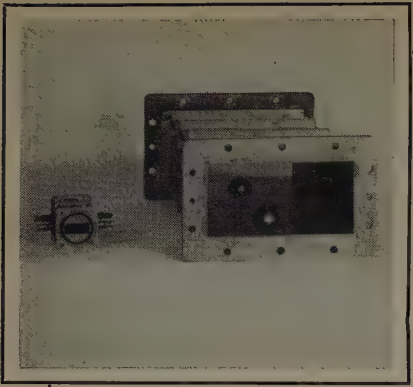

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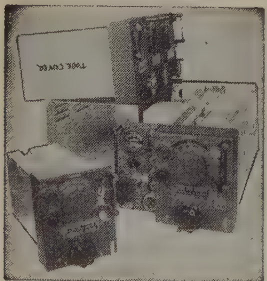

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Langford, R. C., 920 Adams Ave., Elizabeth, N. J.
McCuaig, J. E., c/o Cal-Tech (J.P.L.), White Sands Proving Grounds, N. Mex.
McElroy, E. J., Box 424, Scott AFB, Ill.
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Moos, A. M., Patterson, Moos and Co., Inc., 90-28 Van Wyck Expressway, Jamaica 18, L. I., N. Y.
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Winquist, R. T., 234 West Ave., Bridgeport 3, Conn.
Wright, W. F., 601 Winston Dr., Vestal, N. Y.
Wulfmeyer, J. A., 9428 Brenda, Affton 23, Mo.
Yackey, C. H., 4429 Mariota Ave., N. Hollywood, Calif.
Zettler, S. G., 1518 St. Peters Ave., New York 61, N. Y.
Zimmer, H. W., Sylvania Electric Productis, Inc., 1740 Broadway, New York 19, N. Y.

The following admissions to the Associate grade were approved to be effective as of September 1, 1954:

Abend, D. L., 1057 Southern St., Seattle 8, Wash.
Anderson, C. R., 3070 Grape St., Denver 7, Colo.
Arrowsmith, N. M., 10616 McBroom St., Sunland, Calif.
Baker, D. W., 56 Wigglesworth St., Malden 48, Mass.
Barrow, E. M., 415 Winston Dr., San Francisco 27, Calif.
Baugher, J. D., 5922 Gardendale St., Hollydale, Calif.
Beaulieu, D. E., 230 D Garfield Ave., Collingswood 7, N. J.
BeLusko, J. V., 2712 N.W. Raleigh St., Portland 10, Ore.
Blatchford, D., 22134 Ybarra Rd., Woodland Hills, Calif.
Bohn, E. V., University of British Columbia, 5902 University Blvd., Vancouver, B. C., Canada

(Continued on page 99A)

MEMBERSHIP

(Continued from page 96A)

Boothroyd, J. H., 6444 West 87 Pl., Los Angeles 45, Calif.
Bracken, T. C., 108–26—271 Pl., Queens Village 29, L. I., N. Y.
Brownlee, D. K., 238 Roslyn Rd., Winnipeg 13, Man., Canada
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Daniels, R., Monument St., Concord, Mass.
Davis, C. B., 4160 Meadows Dr., Apt. D-2, Indianapolis, Ind.
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Fernald, A. F., 75 Savannah Hiwy., Charleston, S. C.
Floyd, G. W., 8153 Kenyon Ave., Los Angeles 45, Calif.
Gaudinier, B. J., Box 1108, 3337th T.T. Sqd., Scott AFB, Ill.
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Gustafson, D. W., 9 Mechanic St., Attleboro, Mass.
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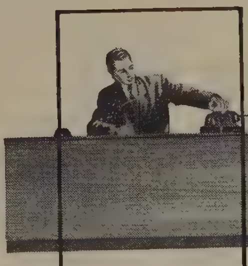
Harris, J. E., Box 536, 3337th TT Sq., Scott AFB, Ill.
Harris, J. E., 22 Berwin St., Johnson City, N. Y.
Hollinger, E. C., 4990 Airway Rd., Dayton 3, Ohio
Holt, T. C., The Superior Oil Co., Midland, Tex.
Irvine, P. L., 677 Fifth Ave., New York 22, N. Y.
Ivanhoe, L., 8A3 Redfield Village, Metuchen, N. J.
Jacobson, T. R., 1828 Andrews Ave., New York 53, N. Y.
Jansson, B. G., 350 Fifth Ave., Suite 7323, New York, N. Y.
Jarett, D., Jr., Fairchild Pl., Whippany, N. J.
Jones, E., Jr., 108 S. First St., Belleville, Ill.
Keene, E. F., 2 Cross St., Kingston, Mass.
Keyser, K. E., 4328 S.E. 47 Ave., Portland 6, Ore.
Khan, M. J. D., Assistant Engineer, Telephones, Cantonment Exchange, Karachi 4, West Pakistan
Koller, E. L., 508 W. Lonnquist Pkwy., Mt. Prospect, Ill.
Korlishin, L., 247 Chestnut Ave., Kingston, Pa.
Koschmann, A. H., 719 Erie St., S.E., Minneapolis 14, Minn.
Kullback, S., 1259 Van Buren St., N.W., Washington 12, D. C.
Kuntz, H. A., Jr., 2908 Patria St., Winston-Salem, S. C.
Leibfried, T. F., Jr., 201 West 85 St., New York 24, N. Y.
Lima, T. F., 61 W. Harwood Ter., Palisades Park, N. J.
Link, A. W.. 3450 N. Lowell Ave., Chicago 41, Ill.
Martel, A. J., 403 Nash, Akron 4, Ohio
Martinez, A., 109 Brow St., Liverpool, N. Y.
McCullough, O. C., 308 N. Oak St., Belleville, Ill.
Mehan, T. O., 614 Merrill St., Park Ridge, Ill.
Monopoli, R. V., Hq. USAFIT, Wright-Patterson AFB, Ohio
Monroy, J. E., International General Electric, Carrera 7 No. 22-80, Bogota Apartado Postal 1055, Colombia

(Continued on page 100A)

GABRIEL L-band components

for your special job . . .

Now, a complete line of large-size microwave components is available from the Gabriel Laboratories. Whatever L-Band components you need . . . stub tuners, power dividers, waveguide components, rotary joints, hybrid junctions, adjustable phase shifters . . . Gabriel can supply them. If you have a special problem, our staff of designers and engineers is ready to meet your most exacting specifications. Extensive testing equipment and complete production facilities operated by highly-skilled personnel insure exceptional electrical performance and mechanical strength.



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The Gabriel Laboratories

THE GABRIEL COMPANY 135 Crescent Street, Needham Heights, Mass.

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The Burlington Meter was chosen for this Bendix-Friez Laboratory Temperature Indicator because they found it met their requirements for an accurate, yet low cost, meter and enabled them to set a desirable price on their instrument. Other famous-name manufacturers have made their selection from the wide ranges, styles and sizes offered by Burlington. Or, let Burlington build a meter to your specifications.

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CUSTOM BUILT TO SPECIFICATION

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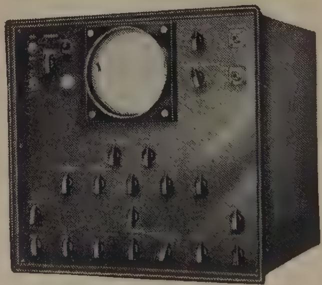
MODEL VDE-4

▶PORTABLE OR RACK MOUNTING

A new, completely self-contained vector display monitoring and test instrument for color. Built-in, high-definition display oscilloscope, with edge-lighted calibrated overlay. Camera mounting facilities for photographic records of vector display, plus continuous operating check. Complete self-checking features for accurate and dependable signal certification. Built-in power supplies. Brackets for optional rack mounting. Complete specifications on request.

Data sheets are available on our complete line of color television instruments for signal generating, distribution, test, and measurement.



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(Continued from page 99A)

Nilsson, W. H., 2373 Palmetto Pl., Savannah, Ga.
Palmer, D. R., 1401 S. Chandler, Denison, Tex.
Parker, W. J., 2282 W. Dodge Rd., Clio, Mich.
Propster, C. H., Jr., Parkway Apts., 50-D, Haddonfield, N. J.
Quinn, T. J., Jr., 36 Lambourne Rd., Towson 4, Md.
Raper, J. A. A., 107 Roney La., Syracuse 5, N. Y.
Rideout, H. A., 720 S. Jackson, Belleville, Ill.
Rinn, E. J. J., 366 Windsor Ave., Windsor, Ont., Canada
Rose, A. C., Bell Telephone Laboratories, Inc., Mountain Ave., Murray Hill, N. J.
Rose, T. L., 105 Pryor, Apt. 534, Scott AFB, Ill.
Rust, D. C., 2919 Kipling St., Palo Alto, Calif.
Sansevero, V. J., 43 Springdale Ave., White Plains' N. Y.
Saums, H. L., Anaconda Wire and Cable Co., Muskegon, Mich.
Schlueter, A. W., 6073 N. Albany Ave., Chicago 45, Ill.
Scull, R. C., 1311—23 St., Newport News, Va.
Shoptaugh, J. R., Jr., 2617 Lapeyrouse St., New Orleans 19, La.
Short, L. E., 3109 St. Clair Ave., East St. Louis, Ill.
Smith, E. J., 492 Orange St., Newark 7, N. J.
Smith, P. G., 60 C Braun Pl., Clifton, N. J.
Stearns, L. D., 2744 Altura Ave., La Crescenta, Calif.
Stewart, R. F., 9017 Pershing Dr., Playa del Rey, Calif.
Taylor, L. S., 34 Summer St., Bedford, Mass.
Thompson, F. T., 302 Glasgow Rd., Pittsburgh 21, Pa.
Town, H. W., 1284 Williams St., Plymouth, Mich.
Tyson, G., 3846 Berryman Ave., Culver City, Calif.
Villavan, Q., 350 N. Garey Ave., Pomona, Calif.
Walley, W. E., 11-C Flannery Ave., Poughkeepsie, N. Y.
Wasserman, J., 179A Lake Ave., Metuchen, N. J.
Weber, J. E., 63 A Farragut St., China Lake, Calif.
Weber, W. E., 4516 Brazil, Los Angeles 39, Calif.
Whalen, P. P., Research Center, Clevite Corp., 540 East 105 St., Cleveland 8, Ohio
White, R. A., Canadian National Steamship, S.S. Canadian Observer, Shed 29, Monteal, Que., Canada
Wines, C. B., Jr., 412 Sixth St., Apt. 5, San Antonio 2, Tex.
Wright, W. E., 1811 Lemoyne Ave., Syracuse, N. Y.
Yeich, V. B., 39 E. Cassily, Springfield, Ohio

Telephone Numbers

of 2,200 Manufacturers

For Your Convenience

are given in the Alphabetical Index to Radio-Electronic Manufacturers in the

New

1954 IRE DIRECTORY

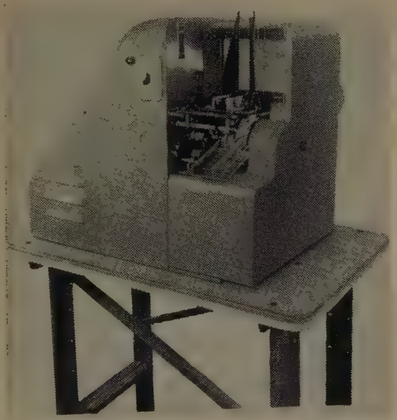
News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 86A)

Color Band Printer

A new semi-automatic color-banding machine, capable of simultaneously printing up to six color coding bands on cylindrical resistors, capacitors and similar objects, is now available from the **Markem Machine Co.,** Keene 42, N. H. Changes in all band colors, or in specific bands only, or from one size object to another, are readily made.



The machine, designated Model 69A, bands components of the following dimensions: $\frac{3}{16}$ to 1 inch in diameter; $\frac{3}{8}$ inch to $2\frac{3}{4}$ inches long (body only); $1\frac{3}{4}$ to 8 inches long over all (wire end to wire end). Both color-band width and space between bands are readily adjustable. Operating speed is governed by the operator's loading speed, which is normally about 50 per minute. Feed from vertical chute to banding position is automatic, as is ejection after printing to container or to conveyor belt (available as accessories).

The machine may also be used for printing complete label detail in one or two colors, on cylindrical objects of the dimensions noted. Complete information is available on request.

(Continued on page 108A)

1955 Radio
Engineering Show
March 21-24, 1955
Kingsbridge Armory
New York

Sel-Rex BRIGHT GOLD PROCESS

NOW USED BY THE NATION'S LARGEST PRODUCER OF AN CONNECTORS

Simplicity of operation . . . uniform, high quality performance . . . characteristics that cannot be achieved with any other gold plating process—just a few of the reasons why American Phenolic Corp. uses the SEL-REX BRIGHT GOLD PROCESS to electroplate high quality AMPHENOL AN Connector contacts.

Here are the advantages of plating with the Sel-Rex BRIGHT GOLD PROCESS:

LONGER LIFE—extremely hard deposit is twice that of conventional gold, thus resists wear and abrasion even on sliding or wiping contacts.

IMPROVED PERFORMANCE—the smooth, fine grained deposit assures a superior contact surface, both electrically and mechanically.

GREATER ECONOMY — uniform metal distribution, even in deep recesses and interior surfaces, gives greater protection with less gold. The solution operates at room temperature and is simple to maintain.

BETTER APPEARANCE — the mirror-like finish will not corrode or oxidize—each part maintains a like-new appearance and has indefinite stock life.

Packaged in 1, 5 and 10 ounce bottles.

SEL-REX PRECIOUS METALS, INC.
Dept. IRE-10, 229 Main Street • Belleville 9, N.J.

COMPACT
DEPENDABLE
EFFICIENT
Rotary Power by Carter

THE NEW Custom DC-AC CONVERTER

These latest of all Carter DC to AC Converters are specially engineered for professional and commercial applications requiring a high capacity source of 60 cycle AC from a DC power supply. Operates from storage batteries, or from DC line voltage. Three "Custom" models, delivering 300, 400, or 500 watts 115 or 220 V. AC. Wide range of input voltage, 12, 24, 32, 64, 110 or 230 V. DC. Unequalled capacity for operating professional recording, sound movie equipment and large screen TV receivers. Available with or without manual frequency control feature.

Dynamotors
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MAIL COUPON FOR CATALOG

Carter Rotary Power Supplies are made in a wide variety of types and capacities for communications, laboratory and industrial applications. Used in aircraft, marine, and mobile radio, geophysical instruments, ignition, timing, etc. MAIL COUPON NOW for complete Dynamotor and Converter Catalogs, with specifications and performance charts on the complete line.

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Please send new catalogs containing complete information on Carter "Custom" Converters and other Rotary Power Supplies.
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...can increase efficiency of your iron core insertion production by 20%

Special embossed construction eliminates torque control problems and stripping . . . prevents breakage or freezing of cores due to cross threading or improper starts.

Custom fabrication to your exact specification assures correct dimensions to within the most critical tolerances, plus uniformity throughout.

Threads are positioned in accordance with your requirement —full thread, each end, one end, center only.

We will furnish—without charge—a pilot production run of custom-made embossed forms to fit your particular application. We will also send a winding mandrel made to the specifications you supply.

Contact us now for full details about this special offer. Request technical bulletin, *Use of Threaded Tubes, Threaded Iron Cores VS. Torque Control.*

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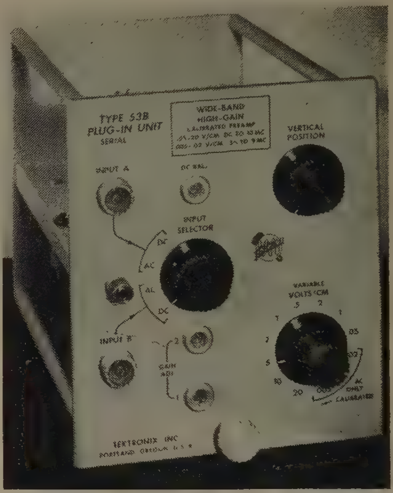
News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 107A)

Wide Band Preamplifier

Tektronix, Inc., P.O. Box 831, Portland 7, Oregon, has developed the Type 53B wide-band high-gain preamplifier, the fourth of a series of plug-in units designed for use with the Tektronix Type 531 and Type 535 oscilloscopes.

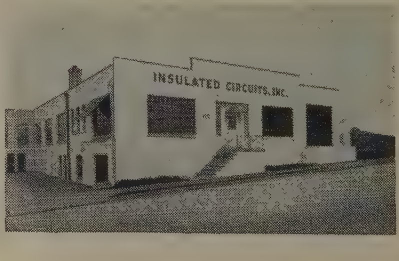


Characteristics provided by Type 53B are: 0.05 v/cm to 50 v/cm sensitivity, ac and dc, continuously variable with nine calibrated steps from 0.05 v/cm to 20 v/cm, dc to 10-mc passband, 0.035-μsec risetime) three additional calibrated sensitivity steps, 5 mv/cm, 10 mv/cm, and 20 mv/cm ac-coupled only, 5-cycle to 9-mc passband, 0.04-μsec risetime; and two input connectors with 80-db isolation. The unit weighs $3\frac{1}{2}$ pounds, and is priced at $125.

New Plant

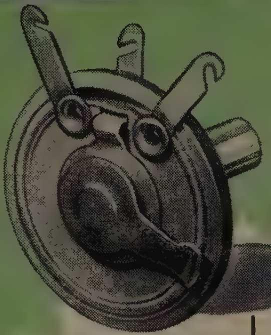
Announcement has been made of the opening of **Insulated Circuits, Inc.,** at 115 Roosevelt Avenue, Belleville, N. J.

The firm manufactures printed circuits and, in addition, their production staff handles complete sub-assemblies.

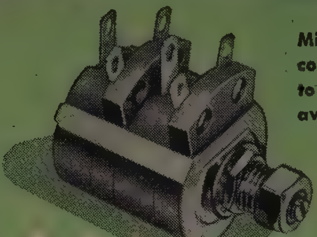


(Continued on page 110A)

Stand Pat with CLAROSTAT

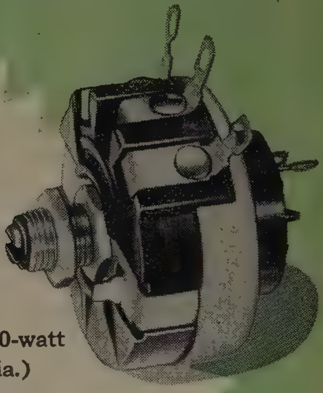


"Humdinger" Series MH ultra-compact potentiometer. 10 to 200,000 ohms. 1 watt.



Miniaturized Series 49 wire-wound controls in single and dual units. 10 to 20,000 ohms. 1.5 watts. Switches available.

"Humdinger" Series 39 shaftless, screwdriver-adjusted potentiometer. 4 to 5000 ohms. 2 watts.



Series 58 3-watt wire-wound controls. 1 to 50,000 ohms. With or without switch.

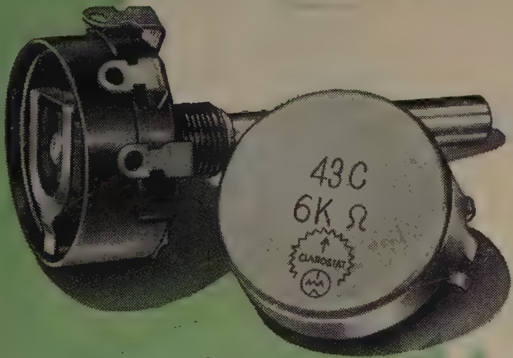
Wire-Wound CONTROLS

Your wire-wound control needs — *usual* or *unusual* — are readily met by specifying CLAROSTAT. Here's why:

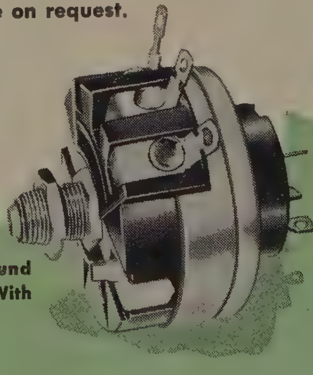
For *usual* needs, the Clarostat line is outstandingly complete. It includes 2-watt (1-1/8" dia.), 3-watt and 4-watt (1-21/32" dia.) types; 25- and 50-watt power rheostats; miniaturized (3/4" dia.) 1.5-watt controls; and the handy, space-saving, cost-reducing "Humdingers"*. All these types, and many more, are *standard and stocked*, available for your convenience at the local Clarostat distributor or in quantities from Clarostat factory stock.

And for *unusual* needs, Clarostat can design and put into production those *special* types — quickly, satisfactorily, economically — often based on ingenious adaptations of standard features and tooling.

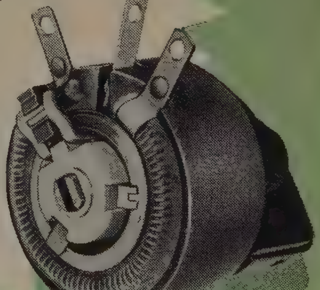
Send those wire-wound control requirements to us for engineering service and quotations. Literature on request.



Smaller or Series 43c 3-watt wire-wound controls. 5 to 10,000 ohms. With or without switch.



Series 10 4-watt wire-wound controls. 1 to 100,000 ohms. With or without switch.



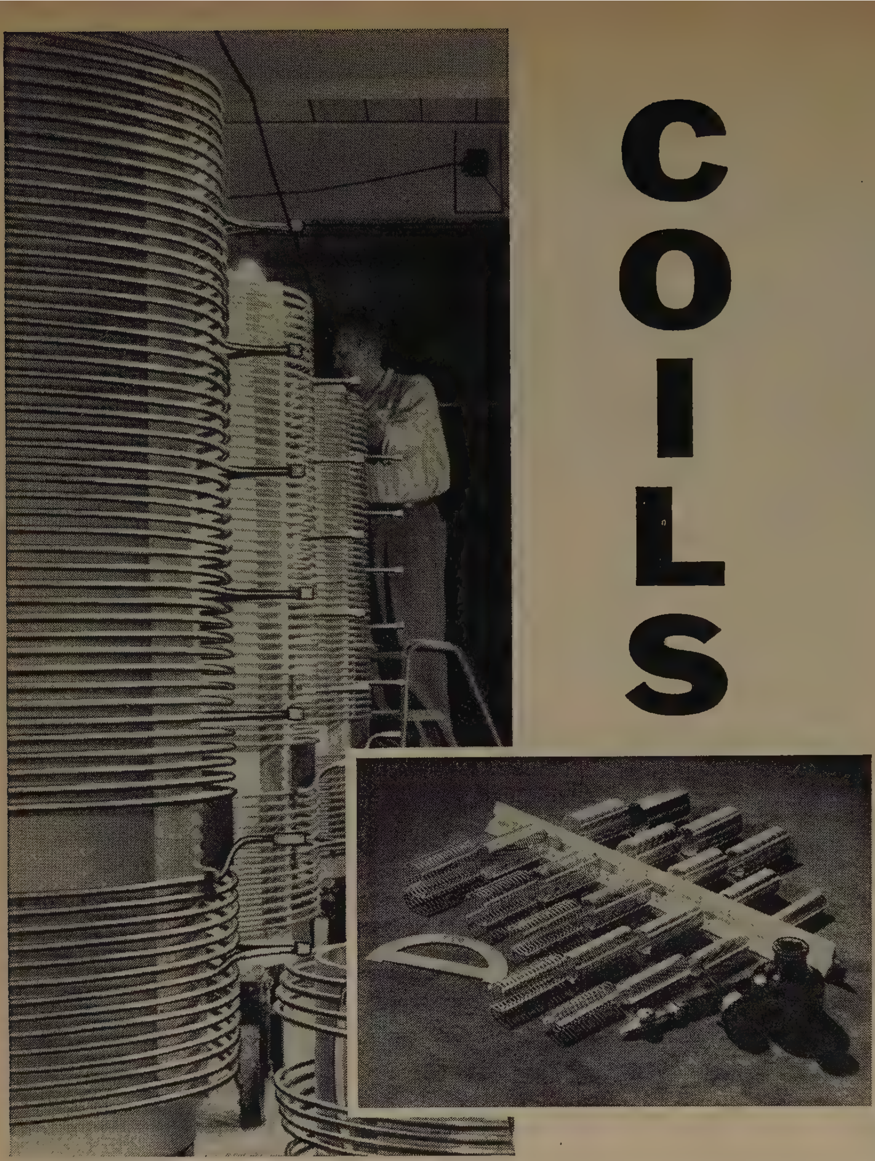
Power rheostats, Series 25 and 50, 25 and 50 watts. 5,000 and 10,000 ohms max., respectively. Also aircraft type, encased in metal housing.

CLAROSTAT®

Controls and Resistors

CLAROSTAT MFG. CO., INC., DOVER, NEW HAMPSHIRE
In Canada: Canadian Marconi, Co., Ltd., Toronto 17, Ont.

*Reg. U.S. Pat. Office



from 6' giants to *half-inch* midgets

When you need coils . . . whether it be just a few for research and development or huge quantities for volume production . . . you'll find your needs right up B&W's alley. Regardless of size or type . . . from 6′ x 48″ variometer type loading coils down to half-inch miniductors, both shown above.

And you'll value the top coil performance you get, based on two decades of specialized manufacture. B&W's twenty years of experience stems from the commercial manufacture of the world's first air-inductor through *single layer solenoid coils . . . universal units with single, multiple pie, or progressive windings . . . r-f, i-f, and oscillator coils . . . traps, discriminators, toroids, filters, r-f and delay line chokes.*

Wherever your coil requirements lie, you'll find Barker & Williamson ready, willing, and able to solve them. Write now for complete information on B&W coil development and manufacturing facilities.

B&W **Barker & Williamson, Inc.**
237 Fairfield Ave., Upper Darby, Pa.

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

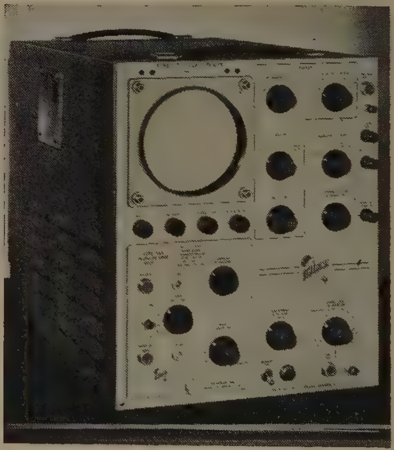
(Continued from page 108A)

Their technique carries the pattern of the printed circuit through the holes to the other side to maintain efficient continuity for faster assemblies, and resulting economy because of the elimination of the usual hardware required.

Quotations are made on design and development, conversion and production. Each job is engineered and a formal quotation is submitted. Large or small quantity orders are accepted.

Oscilloscope

The Type 531, a moderately-priced oscilloscope designed to use plug-in vertical preamplifiers, is available from **Tektronix, Inc.,** P.O. Box 831, Portland 7, Ore. Four calibrated plug-in units are available for use with the Type 531, providing vertical amplifier characteristics for wide-band applications (dc to 10 mc), high dc-sensitivity applications (1 millivolt/cm), and dual-trace applications.



Basic oscilloscope characteristics are: wide-range sweep circuit, 0.1 μsec/cm to 12 sec/cm continuously variable with 24 calibrated steps from 0.1 μsec/cm to 5 sec/cm, accurate 5× magnification on all ranges permitting calibrated sweep times to 0.02 μsec/cm; trigger amplitude selection or automatic triggering; dc-coupled unblanking, 10 kv accelerating potential on new metallized Tektronix crt with helical post-accelerating anode; horizontal input amplifier sensitivity 0.2 v/cm to 20 v/cm continuously variable; 0.25 μsec. vertical signal delay; square-wave amplitude calibrator, 0.2 mv to 100 volts; beam-position indicators show location of the electron beam when it is positioned off the screen.

The unit weighs 61½ pounds, and is priced at $995, plus the price of desired plug-in units.

(Continued on page 112A)

IMPEDANCE BRIDGE
esi

Teamed To Exploit ACCURACY-

MODEL 855-A1

esi OSCILLATOR AMPLIFIER

- Complete operation of the 250-C1 bridge with AC power.
- Highly stable oscillator providing any fixed frequency from 100 CPS to 10 KC with plug-in networks.
- Maximum sensitivity. Visual null indicator driven by highly selective amplifier with over 65 DB gain.

MODEL 250-C1

esi IMPEDANCE BRIDGE

- Most accurate, widest range impedance bridge available. Resistance—1 milliohm to 11 megohms. Capacitance—1 $\mu\mu$f to 1100 μfs. Impedance—1 μh to 1100 henrys.
- Simplified controls. Color index dials. Clearly marked terminals.
- Compact, light, portable. 9″ x 11″ x 11″ over-all.

Featuring **esi DEKADIAL**

Precise resistance, capacitance, inductance readings to four significant figures. Accuracy: resistance ± 0.1%, capacitance ± 0.25%; inductance ± 1.0%.

PRICES

◆ Model 855-A1 Oscillator-Amplifier	$170
◆ Model 250-C1 Impedance Bridge	$340
◆◆ Team	$510

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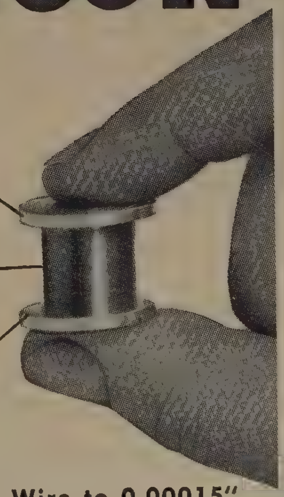
ELECTRO-MEASUREMENTS, INC.

FORMERLY BROWN ELECTRO-MEASUREMENT CORP.

4312 S. E. STARK STREET • PORTLAND 15, OREGON



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round • oval flat grooved ribbon

precision small wire

bare • insulated plated

for highly engineered applications

all metals all alloys

Round Wire to 0.00015″ diameter. Ribbon rolled to 0.0001″ in thickness. Close tolerances held on all specifications.

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... wherever the element calls for PRECISION

development and production metallurgists

SECON METALS CORPORATION

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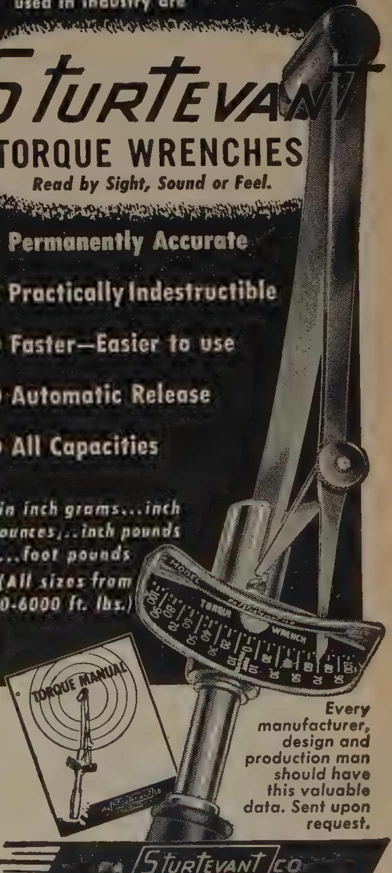
WHite Plains 9-4757

Write for Pamphlet P



Over 85% of the torque wrenches used in industry are

STURTEVANT TORQUE WRENCHES

Read by Sight, Sound or Feel.

- Permanently Accurate
- Practically Indestructible
- Faster—Easier to use
- Automatic Release
- All Capacities

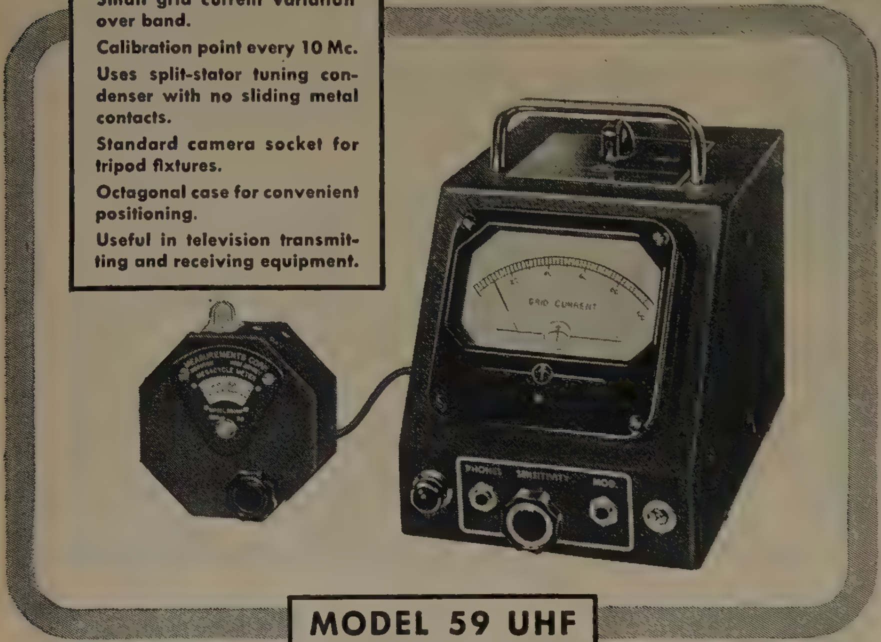
in inch grams...inch ounces...inch pounds ...foot pounds (All sizes from 0-6000 ft. lbs.)

Every manufacturer, design and production man should have this valuable data. Sent upon request.

STURTEVANT CO.
ADDISON QUALITY ILLINOIS

NEW
UHF MEGACYCLE METER
With the Widest Frequency Coverage in a Single Band

FEATURES
- Excellent coupling sensitivity.
- Fixed coupling point.
- Small grid current variation over band.
- Calibration point every 10 Mc.
- Uses split-stator tuning condenser with no sliding metal contacts.
- Standard camera socket for tripod fixtures.
- Octagonal case for convenient positioning.
- Useful in television transmitting and receiving equipment.



MODEL 59 UHF

The Model 59 UHF Megacycle Meter is specially designed to cover the UHF television band plus the many mobile and fixed communication services through its wide frequency coverage of 430 to 940 megacycles in a single range. It incorporates all of the quality features and the excellent performance of the popular Model 59 plus a unique design resulting in a maximum of convenience in use. It is an indispensable instrument for electronic engineers, researchers, experimenters and radio servicemen.

USES
- As a grid-dip oscillator for measuring the resonant frequency of passive circuits such as cavities, tank circuits, inductors, capacitors, chokes, transmission lines and antennas.
- As an auxiliary signal generator for alignment and tuning of UHF receivers and transmitters.
- As an oscillating or absorption marker for use with a sweep-frequency generator.
- As a wavemeter or heterodyne frequency meter.
- As a low sensitivity receiver or field-strength meter for tracing source of spurious oscillations in receivers and transmitters.
- For adjusting antenna systems, wave traps, and filters.

SPECIFICATIONS
FREQUENCY RANGE: 430-940 Mc in a single band
FREQUENCY ACCURACY: ±2% (Individually calibrated)
OUTPUT: CW or 120-cycle modulation
POWER SUPPLY: 117 volts, 60 cycles, 30 watts
DIMENSIONS: Oscillator Unit 4⅜" x 2½"
Power Unit 5⅛" wide x 6⅛" high x 7½" deep

Laboratory Standards
MEASUREMENTS CORPORATION
BOONTON · NEW JERSEY

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 110A)

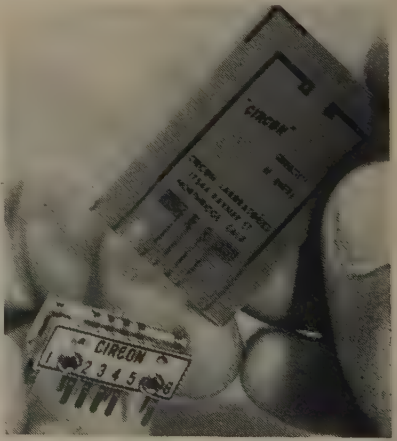
Hibshman Named Secretary of AIEE

The appointment of Nelson S. Hibshman, Brooklyn, N. Y., as Secretary of the **American Institute of Electrical Engineers,** was announced recently by Elgin B. Robertson, President of the Institute. He assumed his new office on May 1.

Mr. Hibshman, who has been Assistant Secretary since January 1, 1953, was named Secretary at a meeting of the Board of Directors in Chicago. He succeeds H. H. Henline of Scarsdale, N. Y.

Printed Circuit Connectors

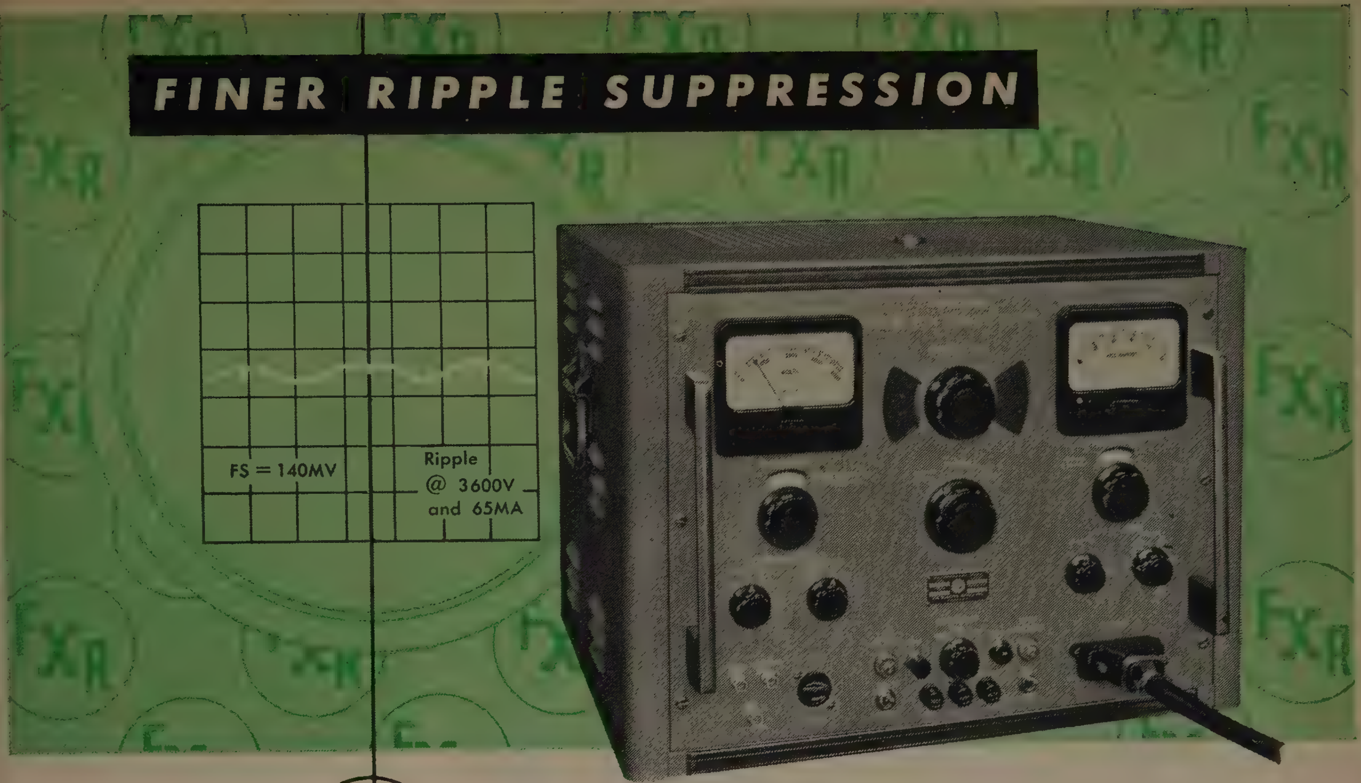
An announcement has been made by the **Circon Component Co.,** 17544 Raymer St., Northridge, Calif., that the new Circon ® miniature and subminiature printed circuit connectors are now in limited production.



The Circon series consists of sixteen models in each of the miniature and subminiature series in basic modules of two,

(Continued on page 114A)

FINER RIPPLE SUPPRESSION



FXR UNIVERSAL KLYSTRON POWER SUPPLY

WIDER RANGE

BEAM:
200 to 2000 volts @ 0 to 100 Ma
1800 to 3600 volts @ 0 to 65 Ma

REFLECTOR:
0 to 1000 volts

CONTROL GRID:
Positive; 0 to 150 volts @ 0 to 5 Ma
Negative; 0 to 300 volts

MODULATOR:
Squarewave; 250 to 2500 CPS @ 0 to 200 volts
Sawtooth; 40 to 120 CPS @ 0 to 200 volts

LOWER RIPPLE

3 MV RMS maximum from any point to ground

GREATER STABILITY

The stability is insured by the use of the 5651 as a reference tube

OPTIMUM REGULATION

All DC voltages regulated to 0.03% for input variation from 105 to 125 volts

Used as a basic tool, the FXR Universal Klystron Power Supply has *proved* its ability for a wide variety of microwave problems. Dependable performance gains its full meaning with the application of this highly versatile, compact unit in the laboratory and for production testing. It is typical of the complete line of FXR Precision Microwave Test Equipment available to science and industry now . . . with the promptness of assembly line production, with the quality of one-of-a-kind design.

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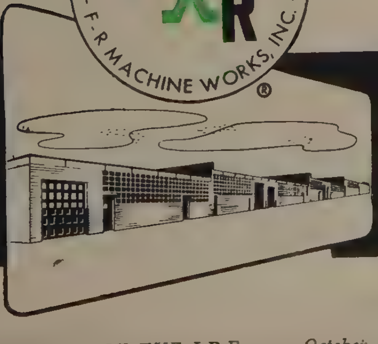
PRECISION MICROWAVE TEST EQUIPMENT

ELECTRONICS & X-RAY DIVISION · FXR · F-R MACHINE WORKS, INC.®



Electronics & X-Ray Division

F-R MACHINE WORKS, Inc.
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TWO
VERSATILE
Electronic Voltmeters

A *Sensitive* VOLTMETER and NULL DETECTOR

Model 310A

AS A VOLTMETER

Frequency Range	10cps—2mc
Voltage Range	100μv—100v
Input Impedance	2meg shunted by 15μμf
Accuracy	3% 10cps—1mc 5% elsewhere

- Voltages as low as 40 microvolts can be measured.

AS A NULL DETECTOR

Frequency Range	5cps—4mc
Threshold Sensitivity	<10μv
Max Scale Sensitivity	10μv/scale division down to 40μv

- Can be used also as wide-band preamplifier with max gain of 60DB and 500 ~ output impedance.

A *Sensitive* VOLTMETER and DECADE AMPLIFIER

(Battery Operated)

Model 302B

AS A VOLTMETER

Frequency Range	2cps—150kc
Voltage Range	100μv—100v
Input Impedance	2meg shunted by 15μμf
Accuracy	3% 5cps—100kc 5% elsewhere

- Ideal for measuring voltages in circuits above ground potential.
- Switch provided for high meter damping.

AS A DECADE AMPLIFIER

Frequency Range	2cps—150kc
Voltage Gains	1000, 100, 10, 1
Output Impedance	approx 3000 ~
Equivalent Input Noise	<10μv

Both Instruments Feature

- Single logarithmic voltage scale with decade range switching.
- Same accuracy of reading at *ALL* points on the scale.

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Write for FREE DB calculator and for complete information on these and other Ballantine instruments.

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102 Fanny Road, Boonton, N.J.

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 112A)

four, six and twelve contacts at rated maximum loads of 1 and 2 amperes per contact. They permit side-by-side and end-to-end mounting of desired combinations and are particularly adaptable to transistor circuitry. Channel design of the female connectors are open-end to accommodate the straight edge of any size of printed circuit without requiring special shapes or fabrication. They will also connect independently to both sides of a circuit and are produced in three gauges to accommodate 1/32, 3/64, and 1/16-inch printed circuits.

The unit is based on an improved principle of contact and connector design. The contact of the female connector is a ribbon of metal, backed with a supporting layer of specially developed elastomer. The compressive load formed by the base upon insertion of the male to the female connector maintains a low resistance contact. Ribbon solder terminals are sufficiently flexible to insert in drilled or punched holes of mating circuits or can be bent or formed as desired.

Beam Power Amplifier

A new beam power amplifier tube, 6094, has been added to the line of reliable tubes produced by **Bendix Aviation Corp., Red Bank Div.,** Eatontown, N. J. This tube is designed to replace the 6AQ5/6005 and other similar tube types.

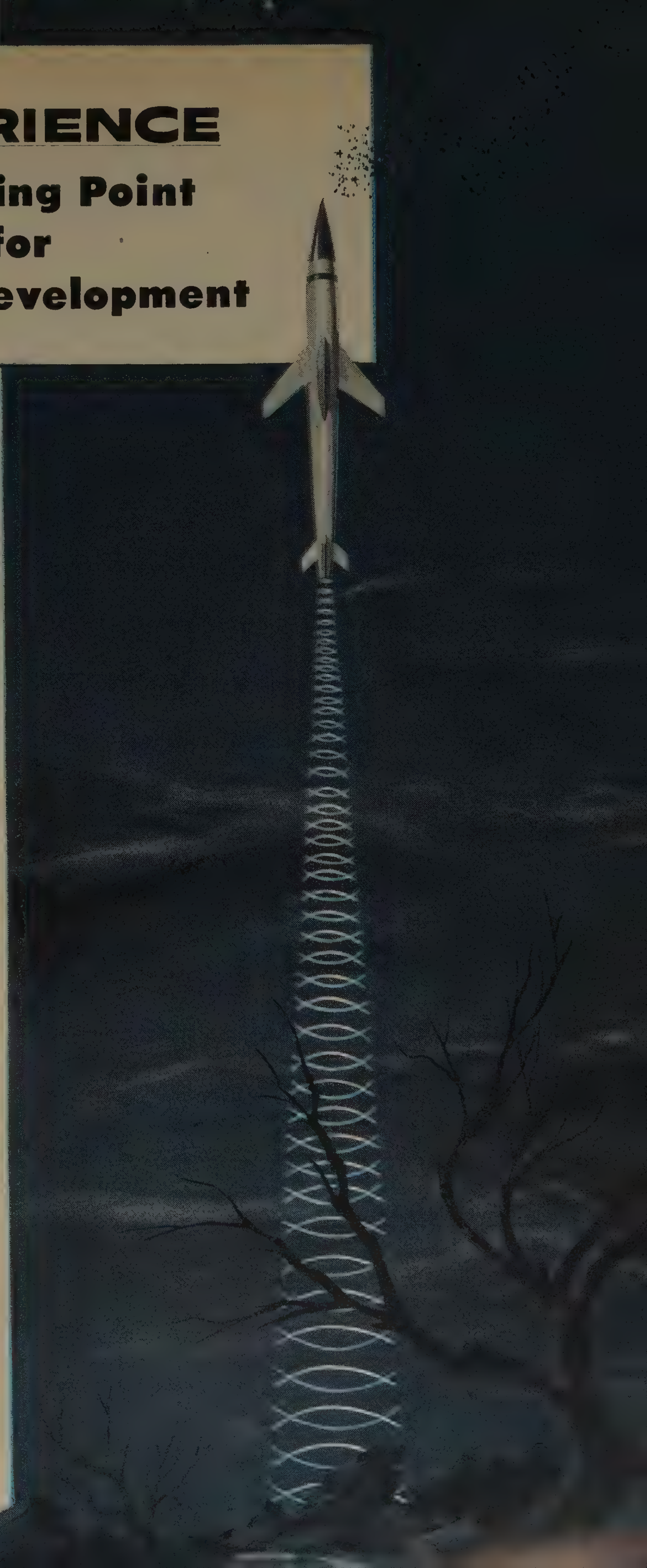
Each 6094 is run-in tested and aged under vibration with all operating voltages applied for forty-five hours. This run-in serves to reduce early failures in actual operation. This tube is designed for use in equipment in which high ambient temperatures and high levels of vibration, shock and other accelerations are encountered. Careful exhaust to a very low vacuum permits operation of the 6094 at bulb temperatures up to 300°C. Ceramic spacers, instead of micas, are used for element separation. These tubes have a cathode-type structure with extruded ceramic heater insulator and an coil type heater, permitting operation at high heater-cathode voltages. The tube has a 9-pin miniature button base and can operate at altitudes up to 80,000 feet.

The 6094 is especially designed for aircraft, industrial, military and other applications where severe environmental conditions are encountered.

(Continued on page 116A)

EXPERIENCE

Launching Point for Missile Development



Ever since 1946, when the Navy flew this nation's irst surface-to-air guided missile, Fairchild has een contributing to the advancement of missile design and development. Fairchild built hat precedent-breaking missile.

The experience gained has been broadened incalculably by the variety of missiles produced for *all* the Armed Services.

Today at Fairchild an integrated engineering team — adept in electronics, air frame structure and aerodynamics, propulsion and in the design of missile ground equipment — is applying the specialized knowledge that only years of experience can bring to a number of current missile projects.

ENGINE AND AIRPLANE CORPORATION

FAIRCHILD

Guided Missiles Division

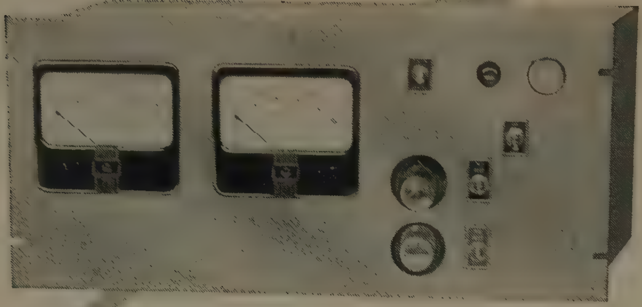
WYANDANCH, N. Y.

•

raft Division, Hagerstown, Maryland • Engine Division, Farmingdale, N. Y.
eed Control Division, St. Augustine, Fla. • Stratos Division, Bay Shore, N. Y.
American Helicopter Division, Manhattan Beach, Calif.

POWER EQUIPMENT COMPANY'S

NEW VOLTAGE REGULATING CIRCUIT



Offers these Design Advantages in Power Supplies!

★ Greater provision for heavier power requirements.

★ Need for fans, blowers or other moving parts eliminated.

★ A VR-105 single voltage regulating tube is only tube used. This has an alternate VR-105 which is used as a ready standby to assure continuous power flow.

★ Filtered to hold ripple voltage in D-C output to less than 0.5% RMS at full load.

★ Not dependent upon accurate maintenance of line frequency. Successfully used with emergency, portable or standby units.

★ New supplies listed in standard sizes.

These new Peco power supplies are designed to do a better job simply, inexpensively and with less maintenance. Write for free bulletin listing specifications and standard sizes.

POWER EQUIPMENT Company

Battery Chargers ☆ Battery Eliminators ☆ D.C. Power Supply Units ☆ Regulated Exciters ☆ and other Special Communications Equipment

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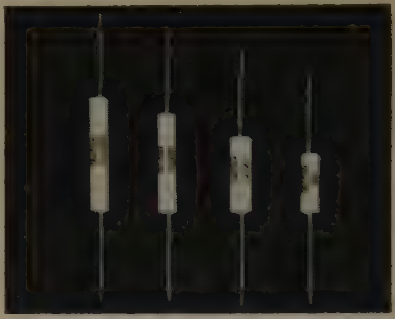
News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 114A)

Axial Lead Resistors

Manufactured by the **TRU-OHM Div., Model Engineering and Manufacturing, Inc.,** 2800 N. Milwaukee Ave., Chicago 18, Illinois, TRU-OHM axial lead resistors are wound on fiber glass cord in a continuous length. The cord is cut to the required length and then the leads are securely clamped to each end. The core is then coated with silicone cement and inserted in a ceramic tube which affords maximum mechanical protection and high dielectric strength. The ends of the resistors are then sealed with silicone cement which precludes any possibility of moisture coming in contact with the resistance element. Axial lead resistors are supplied in standard watt ratings of 5, 7 and 10 watts and to maximum resistance values of 1,000 ohms, 5,000 ohms, and 7,500 ohms respectively.



For more detailed information, write to TRU-OHM.

Digital Computer

The development of a general-purpose digital computer, the Circle Computer, has been announced by **Nuclear Development Associates, Inc.,** 80 Grand St., White Plains, N. Y., its designers, and by **Hogan Laboratories,** 155 Perry St., New York 14, N. Y., the manufacturer. The over all dimensions are 3×4×6 feet.

The Circle Computer contains 700 vacuum tubes and consumes 3½ kw from a 110-volt single-phase power source.

A magnetic drum rotating at 59 rps is employed as the memory element. Its capacity is 1024 words. Each word consists of forty binary digits plus two digits which is equivalent to 12 decimal digits plus two sign digits which is equivalent to 12 decimal digits plus sign. The computation

(Continued on page 118A)

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When Contact Reliability is most important...

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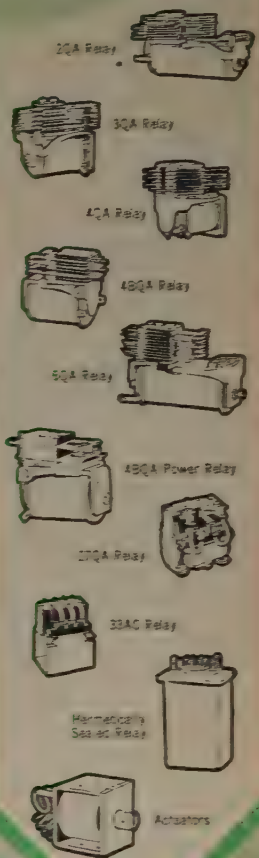
8QA MULTI-CONTACT D.C.

Phil-trol RELAY

...another addition to the many types of Phil-trol Relays...

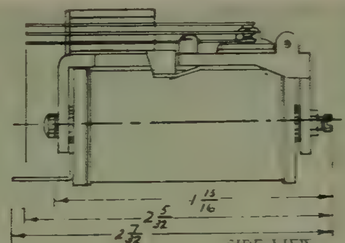


Rugged dependability plus high sensitivity have made this new Phil-trol relay an immediate favorite with engineers and designers. Its compactness and adaptability to a wide range of applications and uses provide new answers to the many complex problems confronting electronic and control engineers.

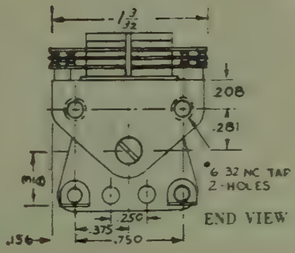
Unusual features of the new 8QA relay include immediate response; fast closing and opening; contact springs with twin contacts; heavy duty, long-life bronze bearings; light weight.

Like all standard Phil-trol relays, the 8QA is available in a wide range of modifications. Coils may be single or double wound, and equipped with copper slugs or sleeves for slow release or for slow operation.

In all probability, there is a standard Phil-trol relay, or variation, to meet your specific need. Phil-trol engineering experience and design facilities are available to help you solve any new application problem.



SIDE VIEW



END VIEW

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PHILLIPS CONTROL CORP.
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OFFICES IN ALL PRINCIPAL CITIES

PHILLIPS CONTROL CORP.,
Dept. PI, Joliet, Illinois

Please send me a free copy of the new Phil-Trol Relay and Actuator Catalog. Also, please arrange to have a Phil-Trol Sales Engineer call on me.

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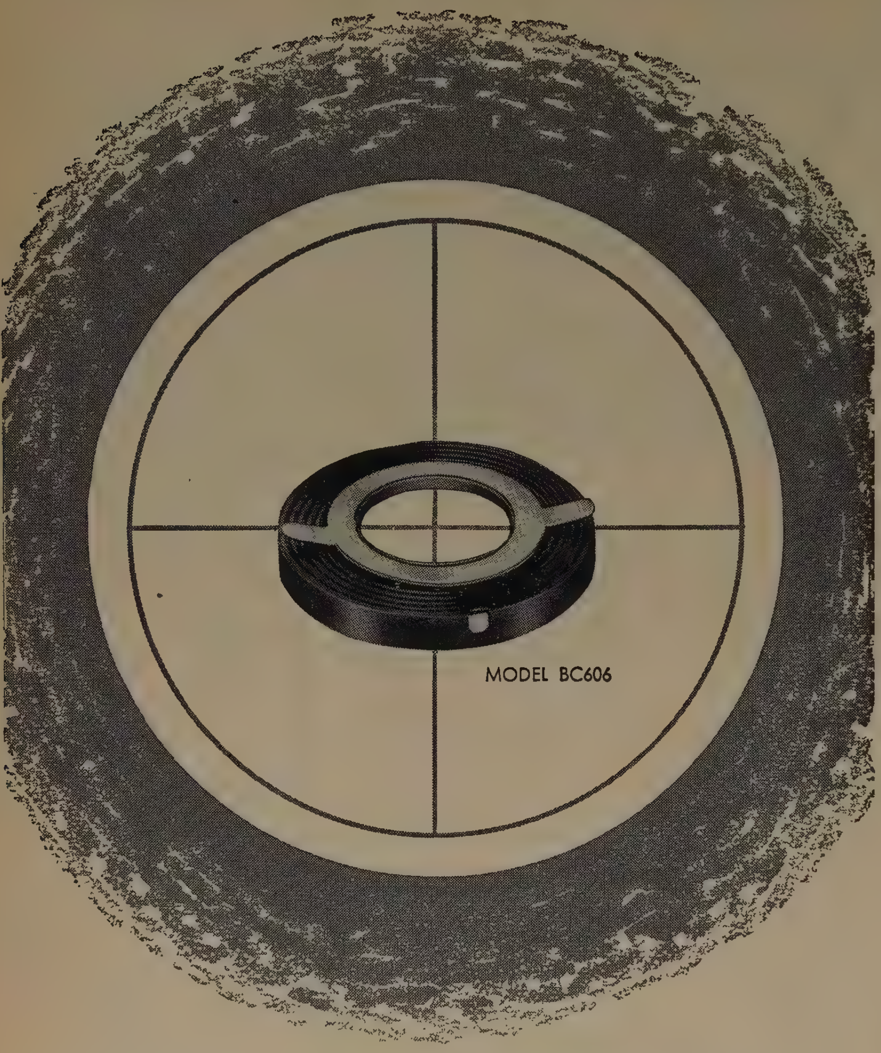
a HEPPNER original!



MODEL BC606

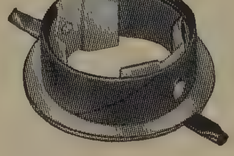
centering devices

For use with Electrostatic TV tubes of all sizes.

Distortion-free beam is assured by uniformity of field. Will not de-focus beam.

The two models differ only in mounting. Model BCC606 mounts easily on the deflection yoke. Model BCC603 mounts directly on the tube, adjacent to the deflection yoke and is held securely in place by phosphor bronze tension springs. Beam centering is done by rotating individual magnets.

Each unit is tested in both open and closed position before shipment.



MODEL BC603

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MANUFACTURING COMPANY
Round Lake, Illinois (50 Miles Northwest of Chicago)
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R. R. 1, U. S. 27, Coldwater Rd.,
Ft. Wayne 3, Indiana
Irv. M. Cochrane Co.
408 So. Alvarado St., Los Angeles, California

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 116A)

speed is indicated by the fact that the multiplication time is 45 milliseconds.

All the controls appear on the operator's console. In addition, neon lights on the control panel allow the operator to monitor the computer's course in a calculation. Input to the computer is by means of an electric typewriter or a punched paper tape reader. For output the computer may operate an electric typewriter, a paper tape punch or both. Intermediate results as well as the final results of a calculation are available at the operator's discretion.

A complete computing installation consists of a Circle Computer with control console and two electric typewriters. The price of the standard model of the Circle Computer is approximately $55,000, exclusive of input and output typewriter. Further information may be obtained from Walter L. Brooks, at the firm.

Traveling-Wave Amplifiers

Two new traveling-wave amplifiers offered by **Hewlett-Packard Co.,** Dept. P., 395 Page Mill Rd., Palo Alto, Calif. provide 30- and 35-db gain over the 2,000- to 4,000-mc band. These amplifiers feature low noise amplification, 1 watt of power output, and millimicrosecond pulsing over a 2,000-mc frequency spectrum. The amplifiers are characterized by a new coupled helix design. In this coupling system there is no mechanical connection between the outer circuitry and the inner tube helix, yet a full transfer of energy is effected. The original and replacement traveling-wave tubes are completely capsulated and adjusted prior to installation in the amplifier.

Two models are available: The Model 490A is used for all types of high gain, broad band, low noise amplification. Its 35 db of amplification is useful with receivers, detectors, and improving the sensitivity of waveguide and coaxial measurement systems. The Model 491A provides a full watt of power output for high level measurements. Thus, it may be connected to a signal generator of 1-mw output to deliver a full watt of power over the 2-kmc to 4-kmc range for antenna measurements, attenuation measurements, impedance measurements, and so forth. Both amplifiers are priced at $1,100.00 f.o.b. factory. Operating details and specifications are available by writing the Hewlett-Packard Co., Dept. P.

(Continued on page 120A)

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---COMPOUNDS

Zophar Waxes, resins and compounds to impregnate, dip, seal, embed, or pot electronic and electrical equipment or components of all types; radio, television, etc. Cold flows from 100°F. to 285°F. Special waxes non-cracking at —76°F. Compounds meeting Government specifications plain or fungus resistant. Let us help you with your engineering problems.

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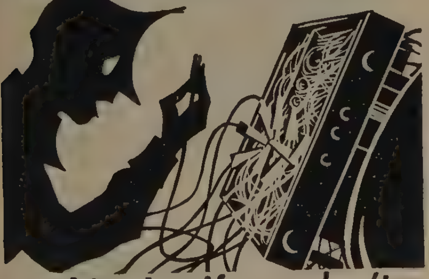

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COMPLETELY INSULATED

Mueller ALLIGATOR CLIPS!

#63 and #63-C insulated Alligator Clip

ASK FACTORY FOR FREE SAMPLE OF #63 CLIP

A skin tight, flexing 2-piece Vinyl insulator covers this clip completely—right down to the nose.

Try the free sample yourself! The absolute utility of these slim, completely insulated clips allows their use in tight spots, where the quick convenience of a clip connection has been impossible in the past. Manufactured as a unit only, with factory-applied insulators.

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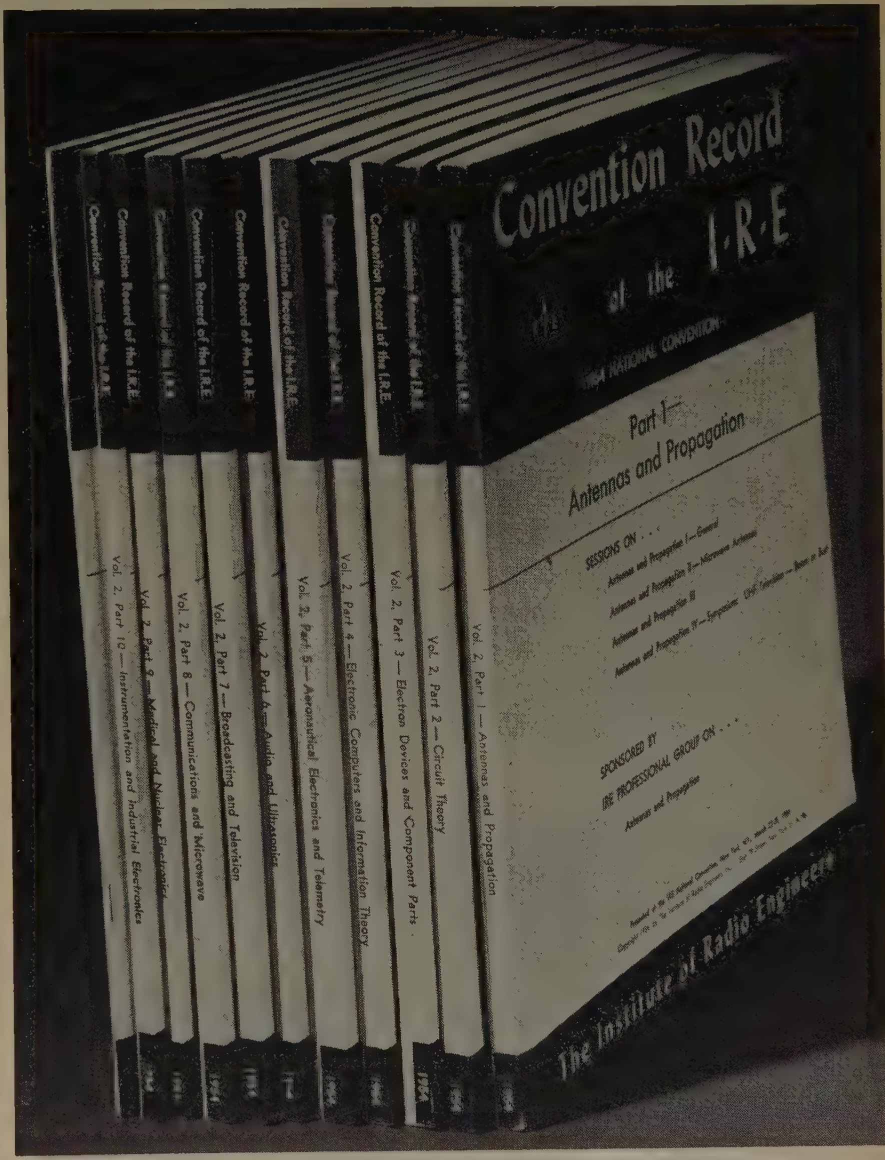
Convention Record of the I·R·E
Part 1— Antennas and Propagation
The Institute of Radio Engineers

In II Parts Organized by Subjects and Sessions

Part	Title	Prices for Members (M) Libraries (L) Non-Members (NM) M	L	NM
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2	Circuit Theory. Sessions: 28, 35, 39, 46	1.25	3.00	3.75
3	Electron Devices & Component Parts. Sessions: 6, 14, 25, 32, 40, 47	1.50	3.60	4.50
4	Electronic Computers & Information Theory. Sessions: 2, 12, 19, 27, 34	1.50	3.60	4.50
5	Aeronautical Electronics & Telemetry. Sessions: 3, 5, 8, 10, 15, 45	1.50	3.60	4.50
6	Audio & Ultrasonics. Sessions: 11, 18, 23, 41, 48	1.50	3.60	4.50
7	Broadcasting & Television. Sessions: 13, 20, 26, 33	1.50	3.60	4.50
8	Communications & Microwave. Sessions: 1, 7, 21, 43, 50	1.50	3.60	4.50
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YOU may buy one or all parts while the editions last. The Convention Record appeared in 10 volumes in 1953 and will be published again in 1955. Please send remittance with order to:

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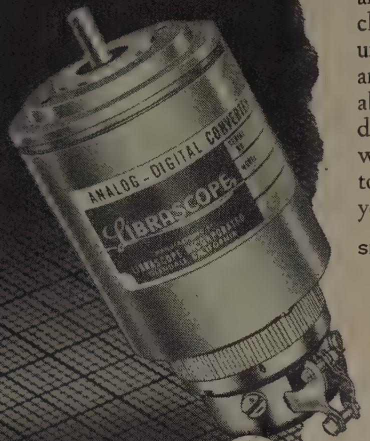
ANALOG-DIGITAL
CONVERTER



NO AMBIGUITY
FAST RESPONSE

A shaft position to digital conversion component for data handling systems and digital computers. Multi-brush pick-offs eliminate ambiguity. Adapted to be used in digital to analog conversion systems and closed loop systems. Standard unit handles 13 digits; 7, 17, and 19 digit models also available. Multi-channel converters designed to be time-shared with one scan network. Custom models may be ordered to your specifications.

SPECIFICATIONS:

Dia. 2 in. Wt. 8 oz.
Length: 2-3/4 to 5", according to model
Torque: Under 0.2 oz. in.
Meets applicable military specifications

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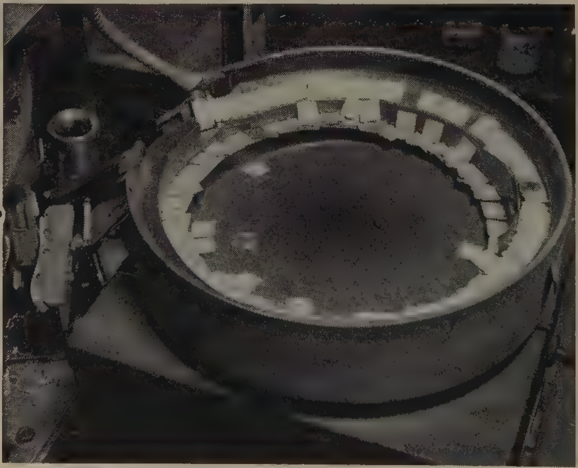
Production
WITH
SYNTRON
PARTS
FEEDERS



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Positive, finger-tip control of feeding rates plus ability to handle small parts of nearly any material or shape enable Syntron Parts Feeders to cut production costs and increase capacities of countless automatic processing installations. Electromagnetic—no mechanical wearing parts—easy to install.

SYNTRON COMPANY
242 Lexington Avenue
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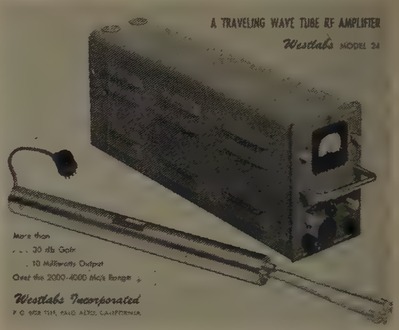
News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 118A)

Broad Band Amplifier

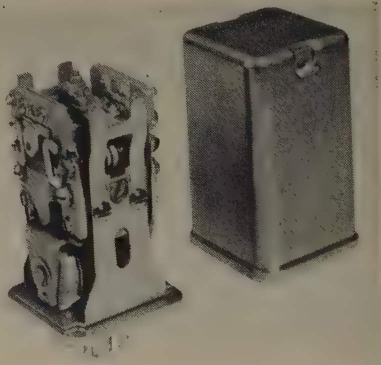
A Model 24 Broad-Band Amplifier which utilizes a traveling wave tube to provide high gain over the 2,000- to 4,000-mc frequency range is announced by **Westlabs Inc.**, P. O. Box 1111, Palo Alto, Calif. The small signal gain averages 35 db, and the saturation output power 15 milliwatts. Maximum noise figure is 20 db or less.


A TRAVELING WAVE TUBE RF AMPLIFIER
Westlabs MODEL 24
More than
. . . 30 db Gain
. 10 Milliwatts Output
Over the 2000-4000 Mc/s Range .
Westlabs Incorporated
P. O. BOX 1111, PALO ALTO, CALIFORNIA

The unit is completely self-contained, including regulated power supplies and traveling wave tube focusing structure. The amplifier is housed in a case of JAN aircraft equipment dimensions ($4\frac{7}{8}$ wide by $7\frac{5}{8}$ high by $19\frac{9}{16}$ inches deep), and is directly usable as either a laboratory tool or a system component. Primary supply requirements are 108 to 122 volts at 1 ampere, 50 to 800 cps.

Servo Relay

Sigma Instruments, Inc., 170 Pearl St., South Braintree, Boston 85, Mass., has developed a new three-position relay which has either a single- or double-pole switch of 2-ampere rating. With current balanced in two windings, or zero in a single winding arrangement, all switch circuits are open. One polarity of coil current or unbalanced current closes one throw of the switch; opposite polarity closes the other. Rated double-pole sensitivity is 12 milliwatts. The relay measures $2\frac{5}{8}$ inches high above octal or magnal socket by $1\frac{5}{8}$ inches square.



Bulletin and application notes on the Type 23JOXCC Relay are available on request from the manufacturer.

(Continued on page 122A)

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Q-max

A-27

LOW-LOSS LACQUER & CEMENT

- Q-Max is widely accepted as the standard for R-F circuit components because it is chemically engineered for this sole purpose.
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- Q-Max is easy to apply, dries quickly and adheres to practically all materials. It is useful over a wide temperature range and serves as a mild flux on tinned surfaces.
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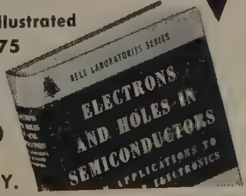
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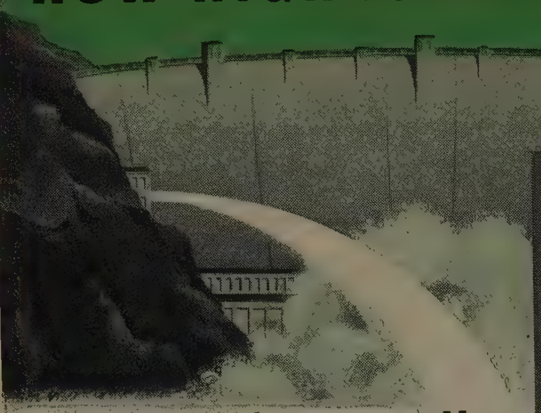
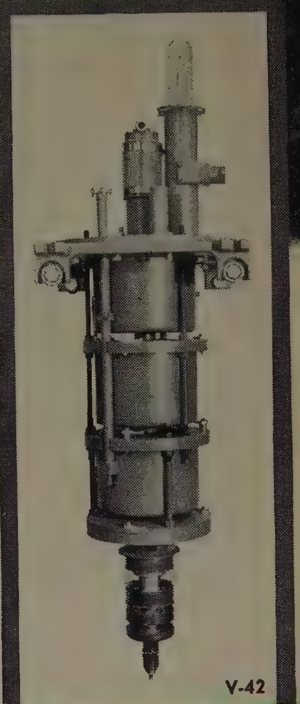
APPLICATIONS TO ELECTRONICS

HOW **HIGH** IS HIGH POWER?




V-42

In klystrons, it's **megawatts** and VARIAN has it . . .

Here are a few of the VARIAN big tubes that answer high power klystron requirements:

FOR: High power microwave communication
UHF-Television transmission

USE THE V-42 SERIES (L-band)

Power output	15 kw CW
Frequency ranges	350 to 1250 mc
Power gain	27 db

FOR: Pulse coherence
Linear accelerators
High power radar transmitters

USE THE VA-80B (S-band)

Pulsed power output	1 megawatt
Power gain	30 db

FOR: Navigation aids
Medium power pulsed systems

USE THE V-82 (X-band)

Pulsed power output	5 kw
Power gain	57 db

HIGH POWER **PLUS**

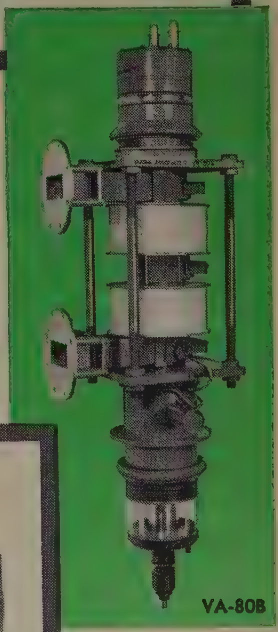
- Unsurpassed frequency stability
- Built-in tuned circuits
- Freedom from maintenance and adjustment
- Reliability and long life

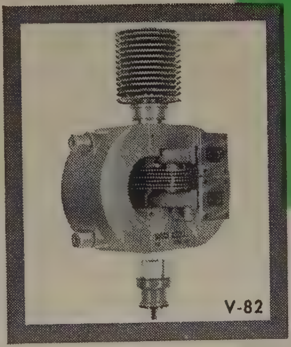
THE **BEST** IN BIG TUBES

These outstanding klystrons exemplify VARIAN design leadership . . . engineering and production skill that consistently delivers quality, economy and unsurpassed performance . . . the reason why VARIAN is the most respected name in klystrons.

FOR COMPLETE SPECIFICATIONS

and application data on these and other VARIAN klystrons, write today to our Application Engineering Department.


VA-80B

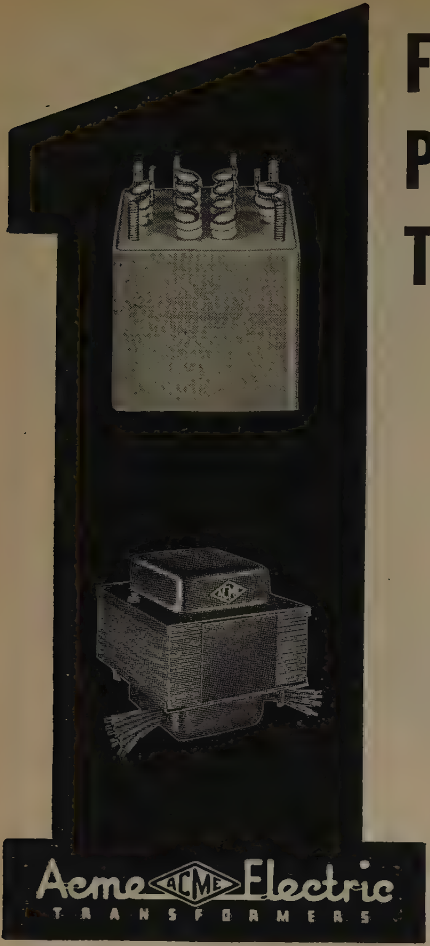

V-82

IN KLYSTRONS, THE MARK OF LEADERSHIP IS

VARIAN associates

PALO ALTO 2, CALIFORNIA

Representatives in all principal cities



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Precision is two-fold in Acme Electric transformers. Exact mechanical dimensions facilitate installation in limited space. Unvarying electrical characteristics provide for dependable performance.

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West Coast Engineering Laboratories:
1375 W. Jefferson Blvd., Los Angeles, California
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50 Northline Road, Toronto, Ontario

for Radar • Communications • Laboratory • Computers

SERIES M200
Miniature i. f. Amplifiers

Here's a brand new way to design i. f. strips! Simply reach for your purchase order pad, specify the characteristics you desire, and mail to Instruments for Industries, Inc. Standard models are available from stock, special designs get prompt attention from our engineering and production staff.

I. F. I.'s fast, convenient service on i. f. strips and other types of broad-band amplifiers has saved valuable design and production time for many leading companies. For complete details write: Instruments for Industries, Inc., 128 Old Country Road, Mineola, N. Y.

INSTRUMENTS ifi INC FOR INDUSTRIES

Instruments for Industries, Inc.

Specifications of Standard Units

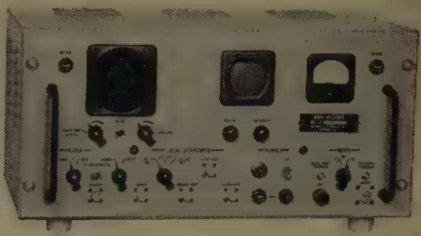
	M260	M230
Band center frequency	60 mc	30 mc
Band width	10 mc	2 mc
Voltage gain	110 db	120 db
Output power	up to 0.02 watts	up to 0.1 watts
Input impedance	50 ohms	50 ohms
Input V. S. W. R.	less than 1.2:1 over pass band	less than 1.2:1 over pass band

Note: M230 model available with 1.5 db noise figure

- 8 — 6AK5's mounted on flat chassis, 1½" x 15"
- Standard BNC cable connectors used for r-f terminals

INSTRUMENTS FOR INDUSTRIES, INC. 128 Old Country Road
Mineola, N.Y.

WWV IS FREE
Why Not Use It?



The WWV frequency standard, as transmitted by the U.S. Bureau of Standards, is the most accurate frequency standard of any in the world.

By using the new Matawan WWV receiver, this frequency standard, with all its information (standard audio frequencies, time signals, propagation notices) is now available to your laboratory. The WWV carrier, after elimination of its modulation, by means of a narrow band pass crystal filter, monitors your local HF standard. The rotative pattern on the screen of the cathode tube gives sense and direction of the frequency deviation.

Either the 5MC WWV or the 15 MC WWV may be received. The receiver will accommodate any local frequency in the range from 50kC to 10MC equal to a subharmonic of 10MC, 2MC or 400 kC. A second channel receives the audio modulation and time signals of the WWV-Standard.

Four models available.

	WWV Input 50 MC	5 and 15.0 MC
Audio Modulation on Speaker Phone Jacks	Model ME-117A $865.00	Model ME-117C $910.00
Additional Filter for 440/600 CPS	Model ME-117B $940.00	Model ME-117D $985.00

F.O.B. Matawan, N.J.

Phillip H. Harrison & Co.
Route 34, Matawan, N.J.
Matawan 1-3113, 1-4242
Write for Illustrated Descriptive Pamphlet

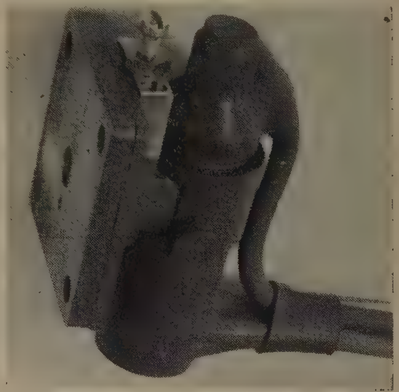
News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 120A)

Klystron for Airborne Applications

Varian Associates, 661 Hansen Way, Palo Alto, Calif., announces a new Ku-band reflex Klystron. The VA-94 is a compact, rugged, low voltage local oscillator type tube, especially designed for missile and radar applications in the frequency range of 16 to 17 kmc. At 300 volts beam potential the VA-94 provides a minimum of 20 milliwatts power output and 55 mc bandwidth, assuring ample performance in mixer and test set applications. Features include the



(Continued on page 124A)

The **RIGHT** *Coated Fabric for* **YOU**

Over the years, leading manufacturers have learned to rely upon APEX when selecting durable, fast colored, sales-improving fabrics for:

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PORTABLE RADIOS
TAPE RECORDERS
TV BOOSTERS
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ANTENNA ROTATOR CONTROL CABLES
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PORTABLE PA SYSTEMS
PORTABLE TV SETS
TEST INSTRUMENTS
SOUND & PROJECTION EQUIPMENT
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Phone: SPring 7-2140

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  760 Bloomfield Ave., Clifton, N.J.
- **CATHODE-RAY TUBES AND COMPONENTS**
  Cathode-ray Tube Division,
  750 Bloomfield Ave., Clifton, N.J.
- **COMMUNICATION PRODUCTS**
  Communication Products Division,
  1500 Main Ave., Clifton, N.J.

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ALLEN B. DU MONT LABORATORIES, INC.

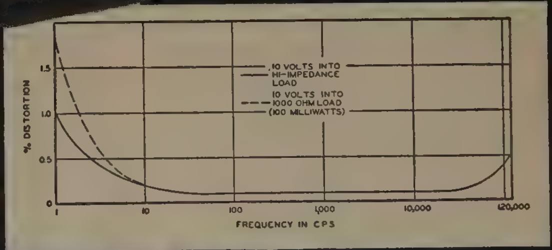
SIE

MODEL M-2 *Low Distortion*

R-C OSCILLATOR

Unexcelled for:

★ Amplifier & filter design
★ Galvanometer manufacture
★ Vibration Analysis
★ Variable frequency standard
★ Transformer & servo design


1.5
1.0
0.5
0
% DISTORTION
10 VOLTS INTO HI-IMPEDANCE LOAD
10 VOLTS INTO 1000 OHM LOAD (100 MILLIWATTS)
1
10
100
1,000
10,000
120,000
FREQUENCY IN CPS

Guaranteed **SPECIFICATIONS:**

Wide frequency range
1 to 120,000 cycles per second
Accurate dial calibration
within 1½% ± .1 cps
Fully regulated power supply
ripple less than .01% of output
Low frequency drift
less than .1% over long term
Excellent amplitude stability
within ½ db. throughout range
High output
20 volts into 1000 ohms or more
12 volts into 600 ohms
1 volt at 300 ohms constant impedance

425⁰⁰

SIE

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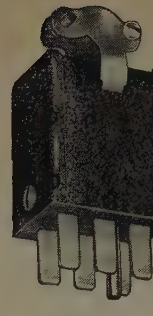
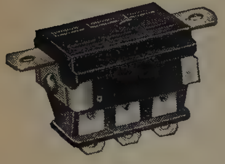


P-406-CCT



S-406-AB

Double Contact Area

Phosphor bronze knife-switch socket contacts engage both sides of flat plug contacts.

Socket contacts phosphor bronze, cadmium plated. Plug contacts hard brass, cadmium plated. Insulation molded bakelite. Plugs and sockets polarized. Steel caps with baked crackle enamel. 2, 4, 6, 8, 10, 12 contacts. Cap or panel mounting.

Information on complete line, in Jones Catalog 20: Electrical Connecting Devices, Plugs, Sockets, Terminal Strips. Write

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News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 122A)

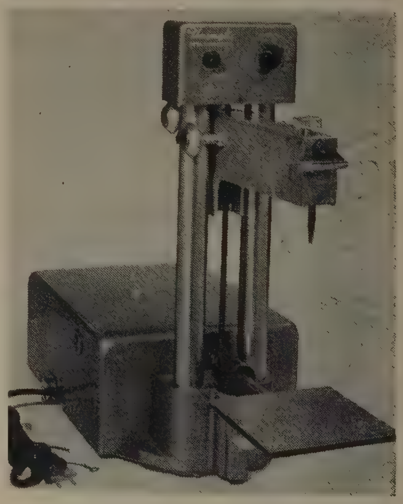
convenience of waveguide output, matched load operation, miniature size (2×1¾×1¾ inches), minimum weight (4 ounces) and Varian's exclusive molded silicone rubber leads for high altitude and high temperature service. A small screw tuner provides convenient frequency adjustment at a slow tuning rate. Microphonics are negligible, even under high amplitude 10G vibration or 150G impact conditions.

Typical Operation

Frequency	16.2	16.7 kmc
Resonator Voltage	300	300 v
Resonator Current	38	38 ma
Reflector Voltage	−145	−155 v
Power Output (VSWR 1.1)	40	40 mw
Electronic Tuning Range	85	65 mc

Spotwelder

The Unimatic Model 1015 Weldmaster, produced by **Unitek Corp.,** 275 N. Halstead Ave., Pasadena 8, Calif., is a precision stored-energy electronic spotwelder intended for electronics and scientific instrument manufacturers. Using the capacitor-discharge principle, the machine has only two stepless controls for the welding variables of electrode force (pressure) and "heat" (energy delivered to the weld).



Model 1015 will weld wires within the diameter range 0.0005-inch through 0.040-inch and sheet metals from 0.0001-inch to 0.020-inch thickness. It produces high strength welds consistently and reliably without discoloration or excessive deformation. The Weldmaster is outstanding in welding metals or combinations of materials usually considered impractical for joining by this method. Among these "specials" are copper, silver, beryllium copper, nickel alloys and many other difficult metals. Parts of widely different thicknesses or gauges are readily welded.

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

Power Supply

A new—hp—Model 712B Power Supply which will provide 0.01 per cent regulation for all conditions of load and power supply variation is available from **Hewlett-Packard Co.,** Dept. P., 395 Page Mill Rd., Palo Alto, Calif. In addition, it has an internal impedance of 0.1 ohm in series with 25 mh, a transient response of less than 1 ms and a hum voltage of less than 500 μv.

This supply has four output options, including a 0- to 500-volt, 200-milliampere regulated supply; a fixed −300-volt regulated supply which may be seriesed with the 500-volt supply providing a 50 ma 500 to 800 volts variable supply for klystron tube operation; a 6.3-volt, 10-ampere unregulated supply; and a continuously variable regulated bias voltage from 0 to 150. There is less than 50-mv change in the regulated supplies from no load to full load and for ±10 per cent line voltage variation.

Other features include separate voltage and current meters and overload protection. The price is $350.00 fob the factory.

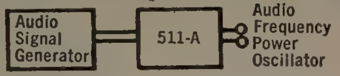
Square Wave Generator

New London Instrument Co., P. O. Box 189, New London, Conn., announces the introduction of a Square Wave Generator to its line of precision electronic measuring equipment.

The new generator, Model 183, provides square waves suitable for testing the transient and frequency response of wide band amplifiers, and accurately measures their amplitude. The frequency range is from 10 cps to 1 mc continuously variable over decade steps.

(Continued on page 135A)

TIE nography
of
INSTRUMENTS



TYPE 511-A POWER AMPLIFIER

- **PHASE SHIFT COMPENSATION**
- **NEGLIGIBLE DISTORTION**
- **HIGH VOLTAGE OUTPUT LEVEL**

. . . a general purpose laboratory power amplifier featuring low distortion, low noise and excellent phase characteristics throughout the frequency range from 50 cps. to 50 kc. A choice of four outputs available to match various loads (5, 25, 200 or 1200 ohms). The 511A Power Amplifier is especially useful as a test driving source for tachometers, synchros, small motors, choppers, electro-mechanical devices and, with an audio frequency signal generator, as a power oscillator.

At rated frequencies and gain settings the overall phase shift is small. A special feature is the phase compensation circuit which permits the overall phase shift to be maintained at a constant value with varying gain. Harmonic distortion and intermodulation distortion are low. Output voltage up to 120 volts into a 1200 ohm load. Operates into loads varying from pure resistance to pure reactance.

The flexible system of phase shift control makes the 511-A Power Amplifier ideal for use in conjunction with phase measuring equipment as a power source in the investigation of phase characteristics of transmission lines, transformers, filters or equalizing networks, saturable reactors, magnetic amplifiers, and in acoustical measurements.

SPECIFICATIONS:

Output Characteristics and Gain (for 0.5% max. allowable harmonic distortion):

OUTPUT SELECTOR (Front Panel Control)	E_{out} Max.	Voltage Gain	Optimum Load	P_{out} Max.
Position 1	8 volts	1.4	5 ohms	12.8 W
Position 2	18 volts	2.8	25 ohms	13.0 W
Position 3	55 volts	8.0	200 ohms	15.1 W
Position 4	120 volts	21.0	1200 ohms	12.0 W

INPUT IMPEDANCE: 100 K ohms shunted by approximately 10 uuf.

FREQUENCY RESPONSE: At 10 watts or less output, essentially flat from 50 cps to 30 kc, down 0.5 db at 50 kc. At 10 to 16 watts, essentially flat from 50 cps to 30 kc, down 1.0 db at 50 kc.

HARMONIC DISTORTION: At 10 watts or less output, less than 0.5% total harmonic distortion (rms). At 10 to 16 watts output, less than 1.0% total harmonic distortion (rms).

PHASE SHIFT: 1.0° ±1.5° from 50 cps to 10 kc.

Phase shift may be compensated at any single frequency to remain constant for all gain settings. Phase shift may also be made zero for a single frequency and a single gain setting.

INTERMODULATION DISTORTION (rms): Less than 0.5% from 50 cps to 15 kc for difference frequency of 150 cycles.

OUTPUT REGULATION: ±5% of rated output voltage from optimum load to open circuit on all ranges.

HUM AND NOISE: Less than 15 mv. with input shorted.

TECHNOLOGY INSTRUMENT CORP.

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West Coast Engineering Facility, P.O. Box 3941, No. Hollywood, Calif., Poplar 5-8620

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WORK ON THE FRONT LINE OF THE NATION'S VITAL DEFENSE PROGRAM. Sandia Corporation is engaged in the development and production of atomic weapons—a challenging new field that offers opportunities in research and development to men with Bachelor's or advanced degrees, with or without applicable experience. Here you can work with able colleagues, eminent consultants and superior facilities on advanced projects of high importance — and also build a permanent career in a rapidly expanding field with a company that recognizes individual ability and initiative.

LIVE IN ALBUQUERQUE, THE HEART OF THE SUNNY SOUTHWEST. Located in the historic Rio Grande Valley at the foot of the Sandia Mountains, mile-high Albuquerque is famous for its climate—mild, dry and sunny the year around. A modern, cosmopolitan city of 150,000, Albuquerque offers unique advantages as a place in which to live. Albuquerque's schools, churches, theaters, parks, and modern shopping facilities afford advantages of metropolitan life—yet hunting, fishing, skiing and a multitude of scenic and historic attractions may all be found within a few hours' drive of the city. New residents have little difficulty in obtaining adequate housing.

ENJOY THESE OTHER IMPORTANT ADVANTAGES. These are permanent positions with Sandia Corporation, a subsidiary of the Western Electric Company, which operates Sandia Laboratory under contract with the Atomic Energy Commission. Working conditions are excellent, and salaries are commensurate with qualifications. Liberal employee benefits include paid vacations, sickness benefits, group life insurance, and a contributory retirement plan. This is not a Civil Service appointment.

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Corporation

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POSITIONS OPEN

The following positions of interest to I.R.E. members have been reported as open. Apply in writing, addressing reply to company mentioned or to Box No. ....

The Institute reserves the right to refuse any announcement without giving a reason for the refusal.

PROCEEDINGS of the I.R.E.
I East 79th St., New York 21, N.Y.

SENIOR ELECTRONIC ENGINEER

Physics or Electronic Engineer from a recognized university with at least 5 years' experience in electronic circuit design. British subject preferred. Salary commensurate with ability. Send resume to Chief Engineer, Cossor (Canada) Ltd., 301 Windsor St., Halifax, N. S. Canada.

ENGINEER

The English Electric Company, Ltd., Luton, England have an interesting vacancy for work on the development of new ideas in microwave circuitry, waveguides and miniaturisation. Apply with full particulars and quoting reference 1160c to Dept. C.P.S., 336/7 Strand, London W.C. 2, England.

ELECTRONIC ENGINEER

An opportunity for young, versatile and ingenious electronic engineers to gain experience in the application of transistors to a wide variety of circuits is offered by the English Electric Company, Ltd., Luton, England. Write with full details to Dept. C.P.S., 336/7 Strand, W.C. 2, London, England, quoting reference 1325.

ELECTRICAL ENGINEER

Electrical Engineer to conduct research pertaining to radar guidance systems. Some knowledge of modern statistics, information theory, radar plan position indicators and their interpretation. Preferably 3 to 5 years' experience. Reply by resume to Personnel Dept., The Franklin Institute, Philadelphia 3, Pa.

TELEVISION BROADCAST TECHNICIANS

CBS has openings in New York City for assistant technicians ($78.00 per week) and full technicians (starting $85.00 per week, progressing automatically in steps to $104.00 after one year and $165.00 after four years) in television studio operations and maintenance. Technical backgrounds and aptitude for operational work are required. Interviews by appointment will be required following written application to CBS television Technical Operations, 524 West 57 St., New York, N.Y.

ENGINEER

Young man wanted with college education, electric and electronic training and sales experience for excellent sales position with one of nation's largest national electrical distributors. Location middle west. Products include broadcasting (AM, FM, TV) equipment, sound, vacuum tube, radio communication and industrial electronic parts. Position offers a conservative, steady, comfortable living backed by sickness, hospital and death benefits, liberal pension plan and company profit-sharing plan. Send summary of qualifications and background to Box 782.

(Continued on page 129A)

Positions Wanted

ENGINEER

BEE 1951, age 29. 3½ years Project Engineer on classified R & D of electronic fire control systems and equipment, radar and infrared; 3 years Radar Officer, U. S. Navy; Varied experience in design and production. 1 year toward LL.B. Post graduate courses in EE. Desires position in R & D, or application. Willing to learn and relocate. Box 791 W.

ENGINEER (No License)

Graduate of Brooklyn High School of Auto. Trades and R.C.A. Institutes. Want position with future; Can work in conjunction with engineering staff in electronics and mechanical fields. Married, age 30. Veteran. Box 792 W.

ENGINEER

BS in Mil. Engineering, MS in Com. Engineering, Ph.D. in Electrical Engineering. Specialized in network theory, circuit design, filter design, transit analysis. Desires association with consulting engineering firm in anticipation of eventual retirement from Government service. Prefer northern New Jersey location. Box 793 W.

ENGINEER

BSEE (Communications) 1952, age 26, married. U. S. Navy Aviation Electronics Technician. Experience in application engineering. Desires position in application or sales engineering. Resume on request. Box 794 W.

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 125A)

The Model 183 has a low impedance output which provides 10 volts p-p. At high impedance 100 volts p-p is available. A 60-db step attenuator and a 20-db continuous attenuator (which do not affect wave shape) provide means of using the generator as a voltage calibrator.

For detailed information, write to the New London Instrument Co.

Portable Tape Recorder

Ampex Corp., Redwood City, Calif., has a new machine, the "600," which will embody professional recording standards in a portable unit.

(Continued on page 142A)

UCLA
Our Patio

OPENING FOR ELECTRONICS ENGINEER AT GILFILLAN

Naturally, no man takes a job in California simply because living is more fun here, or because of fine educational institutions. But when these are added to everything you have been looking for in a lifelong career, including rapid advancement, interest, security and prestige—that's different.

That is what a job with Gilfillan can mean to a qualified engineer; to you, now, because Gilfillan has several openings for good men.

SECURITY

Gilfillan is a major company, with 7 plants located throughout Southern California, that has been in business more than 40 years.

SATISFACTION

Gilfillan is big enough to handle all phases of engineering research, design and production, yet not so big you get lost. You have the very real satisfaction of following through on your work, getting credit for it, and seeing its achievements in the field.

PRESTIGE

Gilfillan pioneered GCA Radar and all its major improvements, is now engaged in automatic GCA, the new lightweight Gilfillan GCA Quadradar, missile guidance, navigational systems, E. C. M.—and a series of classified projects involving advanced and unsolved techniques for all branches of the military services.

Work is interesting at Gilfillan because ideas begin here. As a Gilfillan engineer, your value increases—because of the Gilfillan reputation, and because you work in coming fields.

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You go ahead at Gilfillan as fast as you are ready. You are paid and advanced according to ability, not seniority. If you want further formal training, Gilfillan pays your tuition at UCLA or USC. You can be sure your ability will be recognized, because every man in a supervisory capacity at Gilfillan is a qualified engineer.

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R. E. Bell, Gilfillan Bros., Dept. 104, 1815 Venice Blvd., Los Angeles, California. Give a brief resume of your background and experience. Interview at a convenient location will be arranged for qualified men.

Gilfillan
GCA
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are given in the Alphabetical Index to Radio-Electronic Manufacturers in the

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Electronic and Mechanical Engineers!

Motorola Research Laboratories, located in the healthful climate of Arizona's Valley of the Sun, has several openings for experienced engineers in the following fields:

Electronic research and development for missile guidance, radar and VHF communications.

Mechanical design of missile-borne and vehicular electronics equipment.

Analysis and laboratory work involving development of new types of airborne and ground radiators and waveguide components.

VHF and microwave antenna and waveguide circuitry.

Transistor development

Desire men with B.S. degree or above. Salary commensurate with education and experience. Free health, accident and life insurance. Free hospitalization. Profit sharing. Paid holidays. Sick leave. Vacations. Ideal working conditions. Plenty of housing, reasonably priced. Excellent schools. Exceptionally mild and dry winter climate.

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Motorola Research Laboratory
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TRANSISTORS AND SEMI-CONDUCTOR DEVICES
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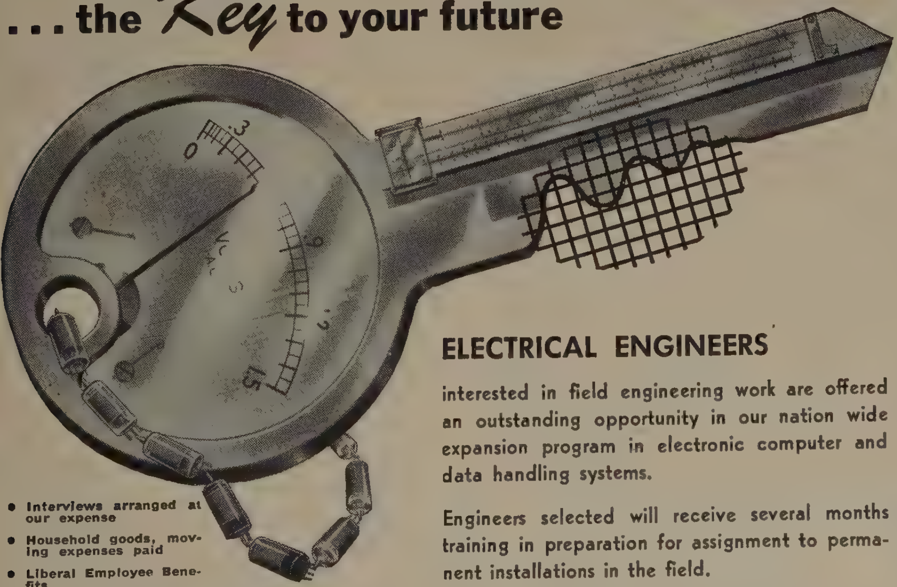

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interested in field engineering work are offered an outstanding opportunity in our nation wide expansion program in electronic computer and data handling systems.

Engineers selected will receive several months training in preparation for assignment to permanent installations in the field.

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Qualified Electrical and Mechanical Engineers and Physicists needed in the following fields:

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Special receivers and transmitters, DF and DME, various instruments and Transistor applications—special devices. Studies in noise, radar, miniaturization and test equipment. Relocating expenses, good insurance plan, central location, steady advancement.

Send resume to L. D. Stearns
Engineering Employment Manager

Hoffman
LABORATORIES, INC.
(SUBSIDIARY OF HOFFMAN RADIO CORP.)
3761 S. HILL ST., LOS ANGELES, CALIF.

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

(Continued from page 135A)

The "600" weighs 28 pounds, in contrast to the 80-odd pounds of previous equipment having the same performance characteristics. Complete with carrying case, the unit measures 16 by 14 inches and is 8 inches thick. The price is $545.00.

The "600" is able to record and reproduce in natural balance the entire audible frequency range from 30 to 15,000 cps. Dynamic range is more than 55 db, and tape speed is 7½ ips.

Air Tool Sets Tab Mounted Controls

The Wire-Wrap Div., Keller Tool Co., Grand Haven, Mich., has just announced its new 14K-1 "Tab-Set" Tool for mounting tab mounted controls to the chassis.

The "Tab-Set," which is air powered, has two spindles that rotate in opposite directions simultaneously, thereby eliminating any reaction to the operator's hand. On the assembly line, the control is inserted into the panel, the "Tab-Set" spindles are placed over the tabs, and when the trigger is squeezed, the tabs are twisted into a locked position. The spindles automatically return to the starting position when the "Tab-Set" Tool is removed.

The light-weight tool, 15 ounces, operates on 60 to 90 psi air pressure. Its price is $68.00, which includes flexible connecting hose and fittings. Request Bulletin 14K-1.

Sine-Cosine Potentiometer

Rawson Electrical Instrument Co., 110 Potter St., Cambridge, Mass., announces production of the RLIICSine-Cosine Potentiometer.

News—New Products

These manufacturers have invited PROCEEDINGS readers to write for literature and further technical information. Please mention your I.R.E. affiliation.

It is useful for slow circular sweep on oscilloscopes or for low-frequency sine-wave generation. The RLIIC provides two output voltages, proportional to the sine and cosine of the angle of shaft rotation. Due to its design and construction of winding and brushes, the RLIIC gives both a sine-wave and a cosine-wave voltage output with a simple linear winding.

It is precision constructed, with ball bearings, precious metal contacts and silver slip rings. The RLIIC has a 360° continuous mechanical rotation, standard winding resistance is 16,000 ohms (± 10) per cent, a rating of 1.5 watts at 65° C., and a life expectancy of 350,000 revolutions, minimum. The output wave is pure sine or cosine, with an average deviation less than 5 per cent. Its over-all diameter is $2\frac{5}{8}$ inches with a case diameter of $2\frac{5}{16}$ inches. Price is $35.00, FOB Cambridge, Mass. Bulletin available.

Connector Catalog

Elco Corp., "M" St. below Erie Ave., Philadelphia 24, Pa., has recently printed a 24 page Varicon Connector Catalog. The complete Varicon line is described as well as an explanation of this miniature Connector system.

Copies of the catalog will be sent, if requested on company letterhead.

(Continued on page 145A)

An Opportunity

is offered for intelligent, imaginative engineers and scientists to join the staff of a progressive and self-sustaining, university-affiliated research and development laboratory. We are desirous of expanding our permanent staff in such fields as electronic instrumentation, missile guidance, microwave applications, design of special-purpose electronic computers, and in various other applied research fields of electronics and physics. Salary structure and benefit programs are on a par with industry. In addition, there are many tangible advantages, such as our self-sponsored internal research policy, of interest to men with ingenuity and initiative.

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Advanced and challenging assignments • Association with top Engineers • Interesting, diversified work • Remuneration based on professional abilities • Modern plant—latest laboratory facilities • Adequate housing • And of course, employee benefits, and congenial surroundings, among the finest.

Please submit resume to our Engineering Personnel Section. (Personal interviews may be arranged in your city.)

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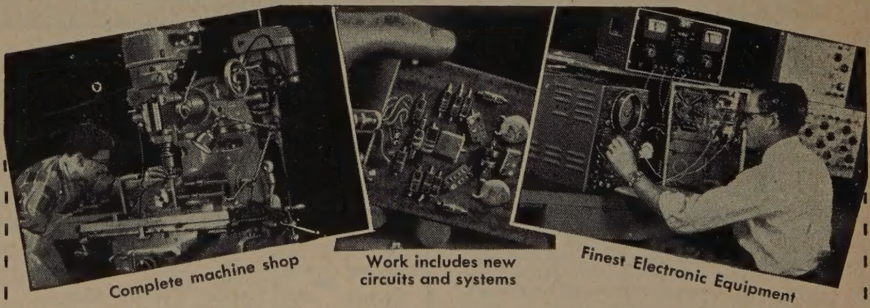
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Work includes new circuits and systems
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Senior engineers to handle design and development projects and provide technical direction of other top-level engineers working on microwave circuits and microwave plumbing in the development of military airborne electronic equipment. Should have 5 years' experience in such work and at least a BS degree.

RADAR SYSTEMS AND CIRCUIT ENGINEER

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INTERVIEWS BY APPOINTMENT
Don Bradley, Personnel Manager, Dept. F, Boston Engineering Lab.

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COMMUNICATIONS EQUIPMENT CO.

MICROWAVE COMPONENTS

10 CM.—RG48/U Waveguide

10CM ECHO BOX: Tunable from 3200-3333 Mc. For checking out radar transmitters, for spectrum analysis, etc. Complete with pickup antenna and coupling devices $27.50

10 CM ANTENNA ASSEMBLY: 3000-3300 Mc. Parabolic Dish, 29 inch Diam. Fed from dipole **Rotation:** 360 Deg. Azimuth at speeds of 20 and 10 RPM. **Tilt:** 30 deg. above and below horizontal. Motor-Driven by 2-28V motors, 4.3 A Total drain. Azimuth info. is fed to selsyn mechanism, and elevation data is obtained from Azimuth potentiometer. Net weight 65 lbs. $78.50

POWER SPLITTER for use with type 726 or any 10 CM Shepherd Klystron. Energy is fed from Klystron antenna through dual pick-up system to 2 type "N" connectors ....$22.50 EACH

LHTR. LIGHTHOUSE ASSEMBLY. Parts of RT39 APG 5 & APG 15. Receiver and Trans. Cavities w/assoc. Tr. Cavity and Type N CPLG. To Recvr. Uses 2C40, 2C43, 1B27. Tunable APX 2400-2700 MCS. Silver Plated $22.50

BEACON LIGHTHOUSE cavity p/o UPN-2 Beacon 10 cm. Mf'g. Bernard Rice, each $27.50

MAGNETRON TO WAVEGUIDE Coupler with 721-A Duplexer Cavity, gold plated $45.00

721A TR BOX complete with tube and tuning plungers $12.50

McNALLY KLYSTRON CAVITIES for 707B or 2K28, 2700-2900 Mc. $4.00

WAVEGUIDE TO ⅞" RIGID COAX "DOOR-KNOB" ADAPTER CHOKE FLANGE SILVER PLATED BROAD BAND $32.50

AS14A AP-10 CM Pick up Dipole with "N" Cables $4.50

HOLMDELL-TO-TYPE "N" Male Adapters, W. E. 2D167284 ....$2.75

I.F. AMP. STRIP: 30 MC. 30 d.b. gain, 4 MC Band-width, uses 6AC7's—with video detector, A.F.C. less tubes $24.00

BEACON ANTENNA. AS31/APN-7 in Lucite Ball, Type "N" feed $22.50

ANTENNA. AT49A/APR: Broadband Conical, 300-3300 MC Type "N" Feed $12.50

"E" PLANE BENDS. 90 deg. less flanges ...$7.50

3 CM.—RG 52/U Waveguide

3 CM ANTENNA ASSEMBLY: Uses 17" paraboloid dish, operating from 24 vdc motor. Beam pattern: 5 deg. in both Azimuth and elevation. Sector Scan: over 160 deg. at 35 scans per minute Elevation Scan: over 2 deg. Tilt: Over 24 deg. $85.00

Cross-Guide Directional Coupler. UG-40 output flange. Main Guide is 6" Long, with 90 Deg. "E" Plane bend at one end, and is fitted with Std. UG 39/UG 40 flanges. Coupling figure: 20 db Nominal $22.50

FLEX. WAVEGUIDE SECTION. 1 ft. long. With UG-40/UG-39 flanges. Attenuation is less than 0.1 db. at 9375 mc. and VSWR is less than 1.02. Rubber covered $7.50

JAN WAVEGUIDE FLANGES

UG 39/U	$1.10	UG 51/U	$1.65
UG 40/U	$1.25	UG 52/U	$3.40
UG 40A/U	$1.85	UG 52A/U	$3.40

VSWR Measuring Section. Consisting of 6" straight section, with 2 pick-up. Type "N" Output Jacks. Mounted ½ Wave apart $7.50

1" x ½" waveguide in 5' lengths. UG 39 flanges to UG40 cover. Silver plated ......**per length** $5.00

Rotating-joints supplied either with or without deck mounting. With UG40 flanges **each.** $17.50

Bulkhead Feed-thru Assembly $15.00

Pressure Gauge Section 15 lb. gauge and press nipple $10.00

Pressure Gauge. 15 lbs. $2.50

Directional Coupler. UG-40/U Take off 20db $17.50

TR-ATR Duplexer section for above $8.50

Rotary joint choke to choke with deck mounting $17.50

90 degree elbows. "E" plane 2½" radius ..$12.50

Microwave Receiver. 3 CM. Sensitivity: 10-13 Watts. Complete with L.O. and AFC Mixer and Waveguide Input Circuits, 6 I.F. Stages give approximately 120 DV gain at a bandwidth of 1.7 MC. Video Bandwidth; 2 MC. Uses latest type AFC circuit. Complete with all tubes, including 723A/B Local Oscillator $175.00

ADAPTER. waveguide to type "N", UG 81/U. p/o TS 12, TS-13, Etc. $14.50

ADAPTER. UG-163/U round cover to special btl. Flange for TS-45, etc. ....$2.50 ea.

1¼" x ⅝" WAVEGUIDE

VSWR SECTION. 6"L. with 2-type "N" pickups mounted ½ wave apart. $7.50

GG 98B/APQ 13 12" Flex. Sect. 1¼" x ⅝" OD $7.50

Slug Tuner Attenuator W.E. guide, gold plated $6.50

BI-Directional Coupler. Type "N" Takeoff 25 db. coupling $22.50

BI-Directional Coupler. UG-52. Takeoff 25 db. coupling $24.95

Waveguide-to-Type "N" Adapter. Broadband $17.50

MAGNETRONS

Type	Peak Range (MC)	Peak Power Out (KW)	Duty Ratio	Price
2J21A	9345-9405	50		$ 9.75
2J22	3267-3333	265		7.50
2J26	2992-3019	275	.002	7.49
2J27	2965-2992	275	.002	19.95
2J29	2914-2939	275	.002	44.95
2J31	2820-2860	285	.002	24.50
2J32	2780-2820	285	.002	39.50
2J38*	3249-3263	5		16.50
2J39*	3267-3333	8.7		24.50
2J48	9310-9320	50	.001	24.50
2J49	9000-9160	50	.001	59.50
2J56*	9215-9275	50	.001	132.50
2J61†	3000-3100	35	.002	34.50
2J62†	2914-3010	35	.002	34.50
3J31	24-27KMC	50	.001	85.00
4J34	2740-2780	900		129.00
4J38	3550-3600	750	.001	169.45
4J42†	670-730	30	.003	169.50
5J23	1044-1056	475	.001	49.00
700B	690-700	40	.002	22.50
700D	710-720	40	.002	39.75
706EY	3038-3069	200	.001	32.50
706CY	2976-3007	200	.001	32.50
725-A	9345-9405	50	.001	Write
QK60†*	2845-3005	.100 CW		85.00
QK61†*	2975-3170	.100 CW		85.00
QK62†*	3135-3330	.100 CW		85.00
QK259†	2700-2900	800	.001	249.00

*—Packaged with magnet.
†—Tunable over indicated range.

KLYSTRONS

723A	$12.50	2K25/723A/B	$27.50
723A/B	19.50	417-A (West'Hse)	17.50

VARISTORS

D-167208	$1.35	D-171812	$1.63
D-171858	$1.42	D-172155	$1.50
D-168687	$1.35	D-167176	$1.25

THERMISTORS

D-164699 Bead Type DCR: 1525-2550 Ohms @ 75 Deg. F. Coefficient: 2% Per. Deg. Fahr. Max. Current 25 MA AC/DC $2.50

D-167332 Bead Type. DCR is 1525-2550 Ohms. Rated 25 MA at .825-1.175 VDC $1.35

D-167613 Disk Type DCR: 335 Ohms @ 75 Deg. F.P.M. 2.5%, 1 Watt $1.35

D-166228 Disk Type 7120 Ohms @ 69°F. 4220 Ohms @ 80°F. 2590 Ohms @ 100°F., 1640 Ohms @ 120°F $1.35

400 CYCLE TRANSFORMERS

(All Primaries 115V. 400 Cycles)

Stock	Ratings	Price
KS9608	1233VCT/.35MA. 1140VCT/.07A	$ 5.79
352-7102	6.3V/2.5A	1.45
M-7472426	1450V/1.0MA. 2.5V/.75A. 6.4V/3.9A. 5V/2A. 6.5V/.3A. P/O ID-39/APG-13	4.95
352-7039	640VCT @ 380MA. 6.3V/.9A. 6.3V6A 5V/6A	5.49
702724	9800/8600 @ 32MA	8.95
KS9584	5000V/290MA. 5V/10A	22.50
KS9607	734VCT/.177A. 1710VCT/.177A	6.79
352-7273	700VCT/350MA. 6.3V0.9A. 6.3V 25.A 6.3V/.08A. 5V/CA	6.95
352-7070	2x2.5V/2.5A (2KV TEST) 6.3V/2.25A. 1200/100/750V. @ .005A	7.45
352-7196	1140/1.25MA. 2.5V/1.75A. 2.5V/1.75A—5KV Test	3.95
352-7176	320VCT/50MA. 4.5V/3A. 6.3VCT/20A. 2x6.3VCT/6A	4.75
RA6400-1	2.5/1.75A. 6.3V/2A—5KV Test	2.39
901692	13V 9A	2.49
901699-501	2.77V @ 4.25A	3.45
901698-501	900V75MA. 100V/.04A	4.29
Ux8855C	900VCT/.067A. 5V/3A	3.79
RA6405-1	800VCT/65MA. 5VCT/3A	3.69
T-48852	700VCT/806MA. 5V/3A. 6V/1.75A	4.25
352-7098	2500V/6MA. 300. VCT. 135MA	5.95
KS 9336	1100V/50MA TAPPED 625V 2.5V/5A	3.95
M-7474319	6.3V/2.7A. 6.3V/.66A. 6.3VCT/21A	4.25
KS8984	27V/4.3A. 6.3/2.9A. 1.25V/.02A	2.95
52C080	650VCT/50MA. 6.3VCT/2A. 5VCT/2A	3.75
32332	400 VCT/35MA. 6.4V/2.5A. 6.4V/.15A	3.85
68G631	1150-0-1150V	2.75
80G198	6VCT/.00006 KVA	1.75
302433A	6.3V/9.1A. 6.3VCT/6.5A. 2.5V/3.5A. 2.5/3.5A	4.85
KS 9445	592VCT/118MA. 6.3V/8.1A. 5V/2A	5.39
KS 9685	6.4/7.5A. 6.4V/3.8A. 6.4/2.5A ALL CT	4.79
70G30G1	600 VCT/36MA	2.65
M-7474318	2100V/.027A	4.95
352-7069	2-2.5V Wdgs at 2.5A. Each Lo-Cap., 22Kv Test	5.95
352-7096	2.5V1.79A. 5V13A. 6.5V/6A. 6.5V/1.2A. D/O BC800	4.95
352-7099	360VCT/20MA. 1500V/1MA. 2.5V/1.75A. 6.3V/2.5A. 6.3V.6A. P/O BC-929	6.45
D163253	5200V-002A. 2.5V/5A	5.35
M-7471957	2.5V/20A. 12KV Test	4.85
352-7179	250V/100MA. 6.5V/12ACT 5V/2A	3.45

PULSE NETWORKS

15A—1-400-50: 15 KV. "A" CKT. 1 microsec. 400 PPS. 50 ohms imp. $24.50

G.E. ‡3E (3-84-810) 8-2.24-405) 50P4T ; 3KV "E" CKT Dual Unit: Unit 1, 3 sections, 0.84 Microsec 810 PPS, 50 ohms imp.; Unit 2, 8 Sections, 2.24 microsec. 405 PPS 50 ohms imp. $6.50

7-5E3-1-200-67P. 7.5 KV "E" Circuit, 1 microsec. 200 PPS, 67 ohms impedance 3 sections $7.50

7-5E4-16-60, 67P, 7.5 KV "E" Circuit, 4 sections 16 microsec, 60 PPS, 67 ohms impedance $15.00

7-5E3-3-200-67P, 7.5 KV, "E" Circuit, 3 microsec, 200 PPS, ohms imp. 3 sections $12.50

‡755: 10KV, 2.2usec., 375 PPS, 50 ohms imp. ...$27.50

‡754: 10KV, 0.85usec., 750 PPS, 50 ohms imp. ..$27.50

KS8865 CHARGING CHOKE: 115-150 H @ .02A, 32—40H @ .08A, 21KV Test $37.50

G.E. 25E5-1-350-50 P2T, "E" SKT, 1 Microsec. Pulse @ 350 PPS, 50 OHMS Impedance $69.50

KS9623 CHARGING CHOKE: 16H @ 75 MA, 380 Ohms DCR, 9000 Vac test $14.95

G.E. 6E3-.5-2000 50 P2T: 6 KV., "E" Circuit 0.5 usec /2000 PPS/50 ohms/2 sections $7.50

PULSE EQUIPMENT

MIT. MOD. 3 HARD TUBE PULSER: Output Pulse Power 144 KW (12 KV at 12 Amp). Duty Ratio .001 max. Pulse duration: 5, 1.0, 2.0 microsec. Input voltage: 115 v 400 to 2400 cps. Uses: 1-715B, 4-829-B 3-72's, 1-73. New ....Less Cover—$135

ASD Modulator Units, mfd. by Sperry. Hard tube pulser delivers Pk. pulse of 144 kw. Similar to Mod. 3 unit. Brand new, less tubes $85.00

Airborne RF head, model AIA, delivers 50 Kw peak output at 9000 mc. at .001 duty. Complete with pulser unit and all tubes. Used, excel. $185.00

PULSE TRANSFORMERS

Westinghouse 4P37: Primary: 50 ohms imp. 750 v. Sec. 15 kv. 1000 ohms imp. Bifilar filament trans. built in. delivers 12.6 v at 2.5 amp. (pri. 115 v. 400 cy.) $37.50

RAYTHEON WX 4298E: Primary 4KV., 1.0 USEC. SEC: 16KV-16 AMP DUTY RATIO: .001 400 CYCLE FIL. TRANS. "BUILT-IN" $42.50

WECO: KS 9948: Primary 700 ohms: Sec: 50 ohms. Plate Voltage: 18 KV. Part of APQ-13 ....$12.50

GE ‡K-2449A
Primary: 9.33 KV, 50 ohms Imp.
Secondary: 28 KV, 450 ohms.
Pulse length: 1.0/5 usec @ 635/120 PPS, Pk Power Out: 1.740 KW
Bifilar: 1.5 amps (as shown) ...$62.50

GE ‡K-2748-A. 0.5 usec @ 2000 Pps. Pk. Pwr. out is 32 KW impedance 40:100 ohm output. Pri. volts 2.3 KV Pk. Sec. volts 11.5 KV Pk. Bifilar rated at 1.3 Amp. Fitted with magnetron well $39.50

K-2745 Primary: 3.1/2.8 KV, 50 ohms Z. Secondary: 14/12.6 KV 1025 ohms Z. Pulse Length: 0.25/1.0 usec @ 600/600 PPS, Pk. Power 200/150 KW. Bifilar: 1.3 Amp. Has "built-in" magnetron well $42.50

K-2461-A. Primary: 3.1/2.6 KV—50 ohms (line). Secondary 14/11.5 KV—1000 ohms Z. Pulse Length: 1 usec @ 600 PPS, Pk. Power Out: 200/150 KW. Bifilar: 1.3 Amp. Fitted with magnetron well .$39.75

K35J45—Pulse Inversion: PRI: 5 KV PK. Pulse Negative. Sec: Pos. Pulse, 4 KV: 1 usec. and .001 DUTY RATIO $6.50

54J318-1—3 wdgs. Ratio: 1:1:1, 1.10 uh. /wdg., 2.5 ohms DCR $3.50

UTAH X-151T-1: Dual Transformer, 2 Wdgs. per section 1:1 Ratio per sec 13 MH inductance 30 ohms DCR $5.00

UTAH X-150T-1: Two sections, 3 Wdgs. per section. 1:1:1 Ratio, 3 MH, 6 ohms DCR per Wdg. $5.00

68G711: Ratio: 4:1 Pri: 200V, Sec. 53V, 1.0 usec Pulse @ 2000 PPS, 0.016 KVA $4.50

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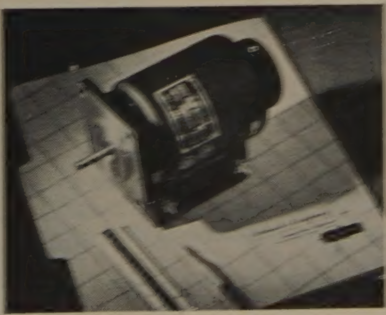

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INDEX AND DISPLAY ADVERTISERS

Meetings with Exhibits 2A
News and New Products 20A
Industrial Engineering Notes 36A
IRE People 78A
Professional Group Meetings 86A
Section Meetings 90A
Student Branch Meetings 90A
Membership 93A
Positions Open 126A
Positions Wanted by Armed Forces Veterans 130A

DISPLAY ADVERTISERS

Acme Electric Corp. 122A
Admiral Corporation 132A
Aerovox Corp. 32A, 33A
Airborne Instruments Laboratory Inc. 56A
Aircraft Armaments, Inc. 132A
Aircraft Radio Corporation 91A
Airtron, Inc. 95A
Allen-Bradley Company 40A
Allied Radio Corp. 124A
American Lava Corp. 17A
American Phenolic Corp. 71A
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Amperex Electronic Corp. 13A
Amperite Company 86A
Apex Coated Fabrics Inc. 123A
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Arnold Engineering Co. 37A
Associated Specialties Co. 104A

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Barker & Williamson, Inc. 110A
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Bendix Aviation Corp., Red Bank Div. 58A
Berkeley, Div. of Beckman Instr. 83A
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Capitol Radio Eng. Institute 104A
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Cascade Research 39A
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Chicago Telephone Supply Corp. 102A, 103A
Clarostat Mfg. Co., Inc. 109A
Cohn Corp., Sigmund 94A
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Cons. Vultee Aircraft Corp. (Convair) 139A
Content, Edward J. 145A
Cornell Aeronautical Lab., Inc. 143A
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Crosley Div. of Avco Mfg. Corp. 142A
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